Proceedings of the I·R·E

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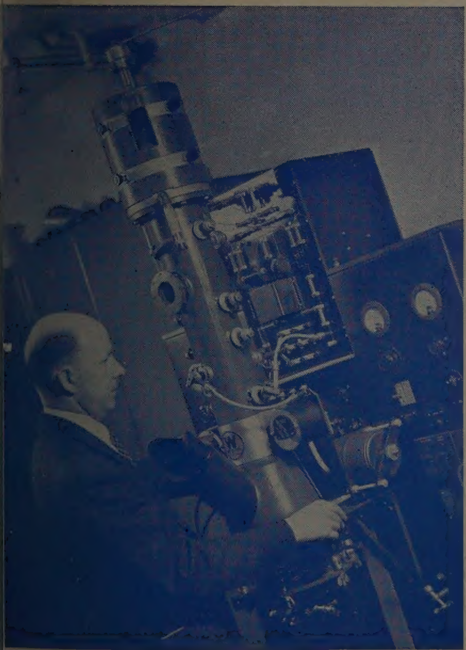


Westinghouse Photo

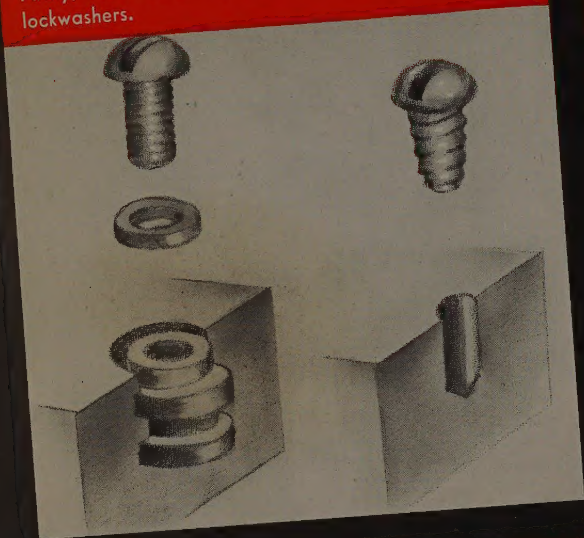
ELECTRONIC WAVE TRACER

Institute of Radio Engineers

In the Battle of Design

A waste of material or machine time in engineering design today is as damnable as sabotage. The battle of design will be won by refinements in existing components as well as by new inventions. Savings in small things add up . . . to big things. Here are some examples:

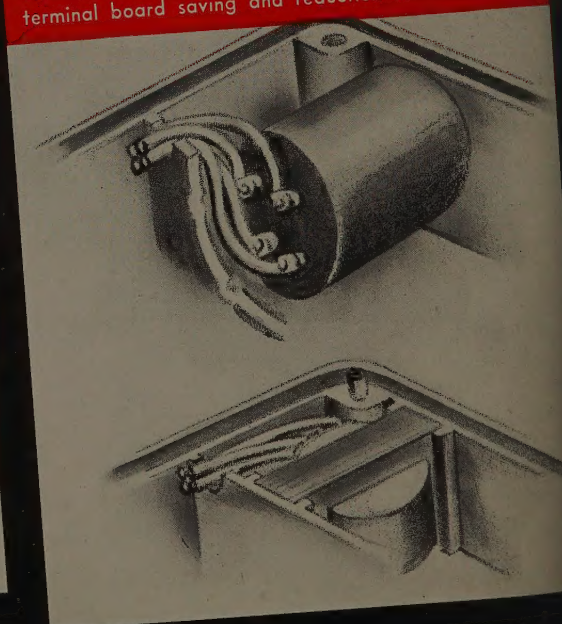
One of our engineers changed the construction of a plastic assembly from brass insert + lockwasher + brass screw to steel PK screw only. Approved by the Army, the savings represented 1,000,000 inserts and lockwashers.



In die cast structures, covers and nameplates were held on by screws. A UTC design modification added a round projection in the casting, which is spun over to hold the plate or cover. Saving: over 2,000,000 screws and lockwashers . . . over 2,000,000 tapping operations.

One UTC design eliminated a threaded shank, lockwasher and nut by changing to a spun-over shoulder on the shank. Saving . . . 150,000 lockwashers and nuts . . . 150,000 threading operations.

This structure employed a cased transformer fastened to a compartment wall with screws. A changed design permitted potting the transformer directly in the compartment. Saving . . . 1,000,000 terminals . . . 300,000 screws . . . 400,000 aluminum cans . . . plus terminal board saving and reduction in overall size.



These savings *added up*. Small in themselves . . . slight for each individual unit . . . their total is impressive. Today we need all possible savings . . . even those which seem impossible at first. Review your designs for Savings for Victory.

UNITED TRANSFORMER CO.

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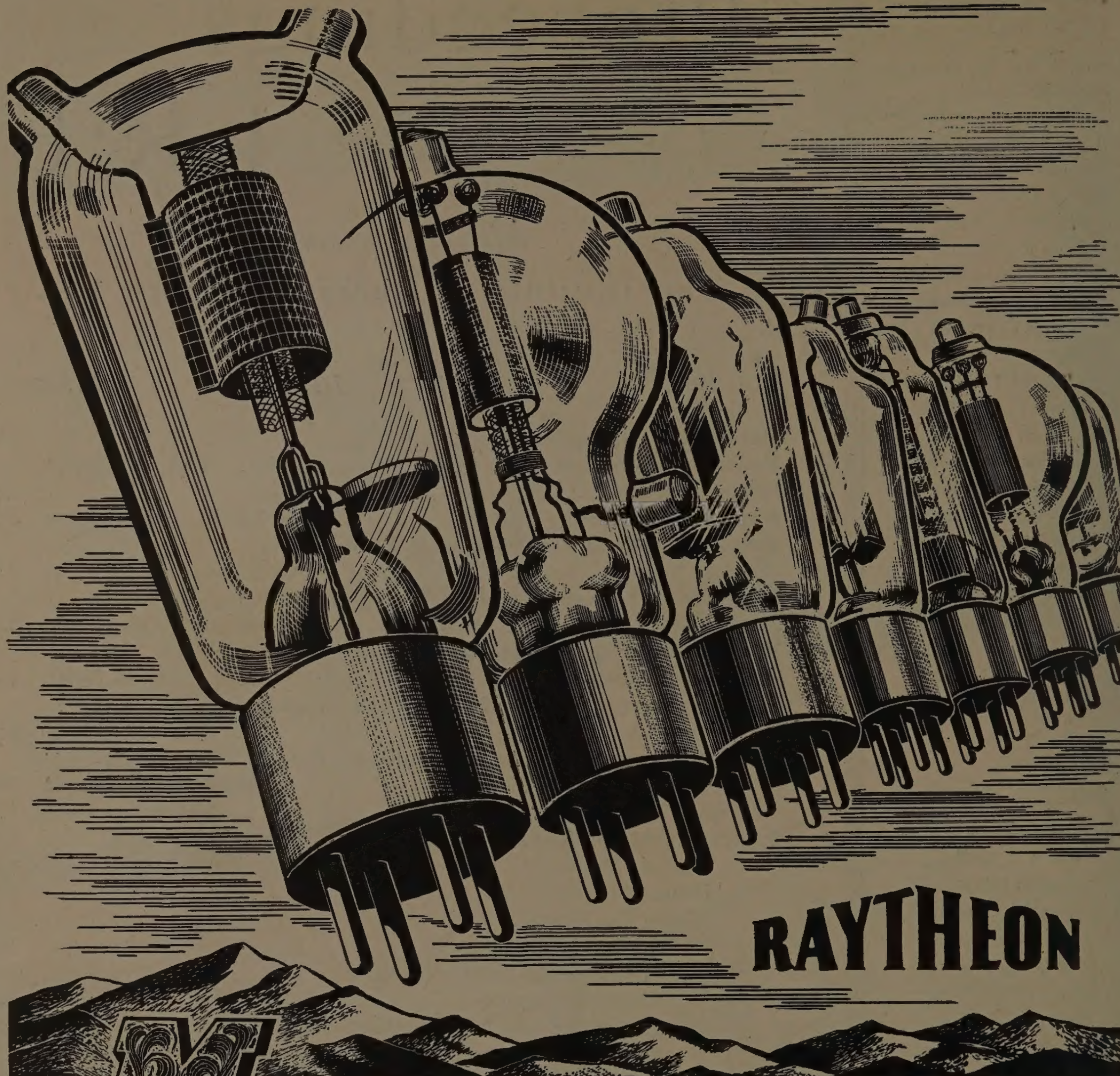

MILEPOSTS IN THE PROGRESS OF RESEARCH!

RAYTHEON tubes for the peacetime electronic era will incorporate all of the engineering skill gained through scientific accomplishments in wartime.

Your new RAYTHEON tubes will be adaptable to a wide scope of newly developed uses . . . with performance characteristics that have been time-tested through service in stringent military campaigns.

You can look to RAYTHEON leadership when you again purchase tubes . . . no matter what type of function your requirements may be . . . you will find a RAYTHEON tube designed and engineered to faithfully perform its task.

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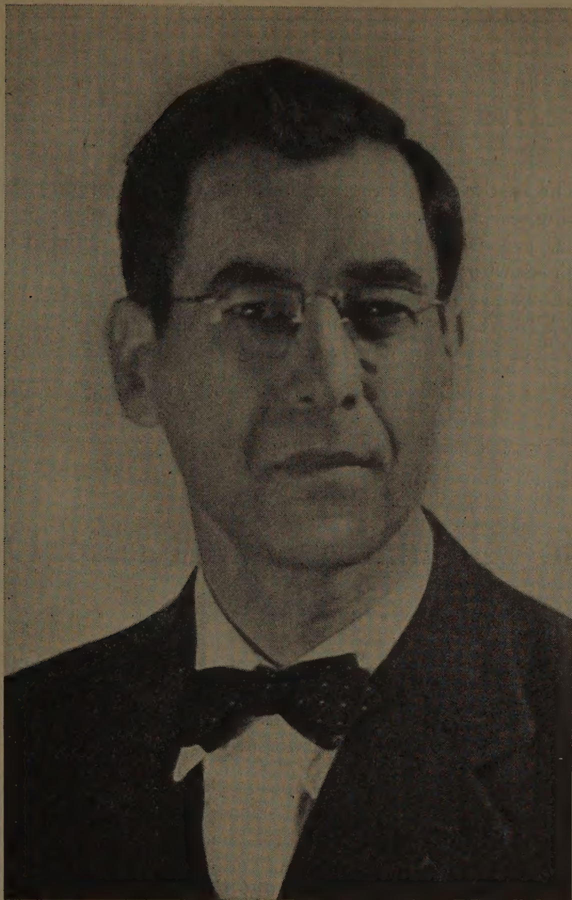
DEVOTED TO RESEARCH AND THE MANUFACTURE OF TUBES AND EQUIPMENT FOR THE NEW ERA OF ELECTRONICS



Harold P. Westman

Secretary, I.R.E. 1930-1942

Harold Prendergast Westman who for over thirteen years was affiliated with The Institute of Radio Engineers, first as Assistant Secretary and then as Secretary, resigned his post on December 15, 1942, to join the staff of the American Standards Association. His new work is an important element of the war effort in that the ASA, working under arrangements with the War Production Board, is preparing a series of standards and specifications for the Military Services. These specifications will be useful to the procurement divisions of the Army and Navy, to radio manufacturers, and to those who supply them with materials and components. The first American War Standard for Fixed Mica-Dielectric Capacitors, prepared by a group under Mr. Westman's chairmanship, has just been published, and other standards on various subjects are being formulated.

Mr. Westman came to the Institute as Assistant Secretary in June, 1929. On the resignation of John M. Clayton as Secretary in February, 1930, Mr. Westman was appointed to that office. He previously had been the technical editor of *QST* and was the author of numerous technical articles for that publication. Previous to his connection with the American Relay League he had worked for the Seigle Laboratories, the Hudson Radio Laboratories, the Radio Audion Company, and the Western Electric Company. He was born in Springfield, Massachusetts, on May 29, 1904. In 1924, he joined the Institute as a Junior member, transferring to Associate in 1925, and to Member grade in 1930. He is a member of the American Radio Relay League, the Veteran Wireless Operators Association, and the Amateur Astronomers Association.

During the depression years, his good judgment, hard work, and lively interest in every phase of the Institute's activity materially helped the organization to come through successfully. His sense of fairness was notable, for he was always able to analyze and understand the viewpoint of others. His many friends in the Institute deeply regret his departure from the Institute but appreciate that he is eminently qualified for his new post and that his work is of vital importance to the war effort of the radio industry. And just as sincerely they wish him success in his future career.

The Institute of Radio Engineers is pleased to present the following searching analysis of past and future radio planning problems, prepared for the PROCEEDINGS by an eminent radio engineer who is a member of the Federal Communications Commission and a Fellow of the Institute of Radio Engineers. Insofar as the Institute and its membership can aid in the solution of the problems which are now faced, they will doubtless endeavor to do so.

The Editor

Radio Regulation and Radio Design

T. A. M. CRAVEN

Not everyone appreciates fully the significance of the relationship between radio-equipment design, the practical operation of radio, and Government regulation of radio. Therefore, a useful purpose may be served by emphasizing a few of the facets of this relationship, as well as the need for co-ordination between these three broad categories of radio.

At the beginning of World War I, the best radio equipment consisted of spark, arc, and high-frequency-alternator transmitters. Reception was accomplished by simple circuit receivers. Some receivers were equipped only with crystal detectors. The best receivers were the so-called autodyne, or self-oscillating type, equipped with two stages of audio amplification. The usable radio-frequency spectrum ranged from about 15 kilocycles to 2000 kilocycles. However, during the war scientific progress was very marked. Radiotelephone transmitters of the vacuum-tube type had been placed in operation and we began to hear of the superheterodyne type of receiver.

Immediately after the war new radio horizons were opened to the public. Broadcasting was started and other new uses of radio began to be applied in public service. The useful radio-frequency spectrum was extended upwards in frequency to about 20,000 kilocycles. Ships and aircraft commenced high-frequency long-distance communication, both by radiotelegraph and by radiotelephone. International point-to-point services were expanded, as well as point-to-point radio services within the United States. Radio began to be used extensively by the Forestry Service, by State and municipal police, and by other Government agencies. The use of radio by amateurs increased tremendously. In the decade from 1920 to 1930 the United States became radio-conscious.

Unfortunately, however, in the small space of seven years, during which the radio spectrum was crowded and unpoliced, chaos existed in the ether. Then in 1927 came radio regulation and international radio-frequency allocation to services, as well as the establishment of the principles of priorities in use of radio frequencies by radio stations capable of causing international interference. However, this radio regulation arrived too late. It was too difficult to unscramble the chaos and start anew on a sound scientific basis. Engineering designs of equipment had been completed and heavy investments had been made in the radio apparatus then in use. The best that could be accomplished by regulation was a compromise between practicalities and science. Many services had to use portions of the radio-frequency spectrum which were not the most useful for specific service. Other services had to share the same portions of the radio-frequency spectrum with different types of services, resulting in confusion and difficulty in the control of interference.

Many wonder how much more valuable radio could have been today if, in 1920, we could have co-ordinated radio design with radio-frequency allocation, or in other words, secured in 1920 comprehensive and co-ordinated general planning by the design engineers, the users of radio, and a Government radio regulatory agency.

We cannot blame anyone for lack of foresight in 1920. We cannot blame radio design engineers for designing equipment for various kinds of radio services using exactly the same portion of the radio-frequency spectrum. However, as late as 1930 we had not learned the lessons of 1920. The radio-frequency spectrum in the decade between 1920 and 1930 had been extended to the ultra-high frequencies, and again, even after three years of Government radio regulation, we discovered that, as the result of the lack of co-ordination between the Government radio regulatory authority and the various elements of the radio industry, radio was fast moving into another chaotic situation in the newly developed portion of the radio-frequency spectrum. We found the equipment used by municipal police, by various Government Departments, including the Army and the Navy, by marine radio services, and by the experimenters in ultra-high-frequency broadcasting, was designed for exactly the same portion of the radio-frequency spectrum. The potentialities for uncontrolled interference threatened the very usefulness of the newly developed radio-frequency spectrum. When the time came to unscramble this chaotic situation, it was discovered that municipalities could not secure additional appropriations to change equipment; the Army and Navy already had a large investment in such equipment. In fact, the same problem existed in 1930 in the then newly discovered portion of the radio-frequency spectrum as existed in 1920. Yet no plans were made, no co-ordination attempted, and the problem remained unsolved until 1936, when compromises were again attempted between practicalities of invested capital and science.

Tomorrow another golden opportunity will be presented to the radio industry. An entirely new radio horizon will be opened to the public. Progress in radio development in World War II has been of tremendous significance, and as a result new uses of radio will be available after the war. Again there will be greater demands than ever before for space in the ether. Once more there is before us the question of whether in radio we shall be unprepared for the peace following a war in which radio science has progressed by leaps and bounds.

We shall have learned much of radio-electronic development ere this war ends. Hence, we could be in a better position to plan for the future after World War II than we were after World War I. Let us not repeat the errors of the past. Let us resolve now to co-ordinate our planning before embarking on a wild scramble of equipment manufacture for the use of new radio channels. Let us avoid in the radio-electronic industry the destruction of radio communications by haphazard design of electronic apparatus not used for communications. Also, let us design radio-communication apparatus for various types of services according to a logical use of the radio-frequency spectrum.

Radio-frequency channels will be at a precious premium, even though the usable portion of the radio-frequency spectrum will be extended into the thousands of megacycles. Therefore, let us plan upon the most scientific allocation of frequencies to the various radio services that we can visualize. Such frequency allocation could easily conform to the practicalities of equipment design as well as to the practical operating conditions in the services where the equipment will be used. Likewise, the radio engineer can attempt to design equipment which will conserve ether space.

The broad phases of engineering design and practical operation can be co-ordinated with a scientific frequency allocation. A long step forward can be taken in the progress of radio. In so doing, we can avoid the pitfalls of premature standardization and its consequent regimentation of research. Thus it seems apparent that as soon as the war clouds show signs of clearing, the entire radio-electronic industry, the radio designer, and the Federal Government should collaborate on the best ways and means to foster the future development of radio on a basis which will minimize, if not eliminate entirely, the potentialities of chaos in the ether of the future.

THE INSTITUTE OF RADIO ENGINEERS
INCORPORATED

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ATLANTA	CHICAGO	CLEVELAND	DETROIT
April 16	April 16	April 22	April 16

LOS ANGELES	NEW YORK	PHILADELPHIA	PITTSBURGH	WASHINGTON
April 20	May 5	May 13	May 10	May 10

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ATLANTA—Chairman, J. M. Comer; Secretary, Ivan Miles, 554—14 St., N. W., Atlanta, Ga.

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BOSTON—Chairman, R. F. Field; Secretary, R. O. Oberg, Northeastern Univ., Boston, Mass.

BUENOS AIRES—Chairman, J. P. Arnaud; Secretary, Alexander Nadosy, Florida St. 1065, Dept. C-16, Buenos Aires, Argentina.

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WASHINGTON—Chairman, C. M. Hunt; Secretary, H. A. Burroughs, Rm. 7207, Federal Communications Commission, Washington, D. C.

Radio Progress During 1942*

Introduction

WAR requirements dominated radio during 1942. During the year there was a great increase in the number of people working on various aspects of radio engineering and radio communication. Manufacturing organizations have greatly increased their staffs engaged in radio engineering and production work. Many schools have been established by civilian and military organizations for training radio technicians and radio operators. A large number of physicists have turned their attention to radio problems. Following the closing of amateur radio stations, substantial numbers of people formerly making radio a hobby have turned to commercial or military radio activities.

As a result of the pressure of military problems, there has been a substantial shortening of the time gap between proved laboratory research results and the stage where mass production of radio equipment can be undertaken.

(1) Frank B. Jewett, "The mobilization of science in national defense," Proc. I.R.E., vol. 30, pp. 113–118; March, 1942.

An interesting aspect of radio during 1942 was the large increase in the number of women engaged in various parts of this field. At the end of the year they were employed as radio research technicians and engineering assistants; as draftsmen, machine attendants, and inspectors in radio factories; as operators in radio stations, and as instructors in radio schools.

A striking change during the year has been the diversion of radio manufacturing facilities from peacetime radio production, generally of a low-precision type, to complex and high-precision war apparatus. The volume of war production of radio equipment during the year was many times the normal annual peacetime output. This increased production was accomplished in spite of the fact that alternatives and substitutes have been required for many materials of which the supply became limited. Purchasing organizations have co-operated with manufacturing establishments by making necessary changes in specifications.

One of the notable fields of development during the year has been the application of radio-frequency principles to many uses other than radio communication. Most of these applications must remain undisclosed for the time being but their fields vary from electrochemistry and metallurgy to biological sciences and agriculture.

Many manufacturers undertook the production of radio equipment for the first time in 1942, adapting their facilities and personnel to the production of radio parts and accessories in close co-operation with organizations responsible for the assembly of complete units. For example, camera factories began making variable condensers, vacuum-cleaner plants began building dynamotors, pen factories stamped out capacitor plates, and watch factories undertook the production of vacuum-tube parts.

As a result of the universal use of radio equipment in modern warfare, the electron tube—perhaps radio's most vital element—is one of the most important single elements in the operations of the great armies and fleets engaged in operations all over the world. The electron tube is used in directing the course of battle, giving running accounts of action to commanding officers and crews, directing gunfire, and determining ocean depth. In aircraft the electron tube assists the maneuvering of planes, and makes possible communication between planes and between plane and ground. Voice communication involving electron tubes carries messages to maneuvering tanks, officers' cars, and even soldiers on foot.

In industrial plants, devices using electron tubes count passing articles faster than the eye can see, automatic sorters discard defective articles, and electronic-control equipments regulate the temperature and other factors, thus aiding in producing greater output of better quality. Among the applications of radio devices which became more widespread during the war are those involving the production of heat in special situations. Thus it may be said that radio waves are now used to weld, rivet, and bake.

In order to simplify the problem of producing electron tubes, an original list of 710 types has been reduced through co-operative engineering study to a little over 100 standard forms of which manufacture is being continued. Interchangeability in manufacture and in service has been facilitated in the United States by the work of a special War Committee on Radio which chose the fixed mica-dielectric capacitator as its initial project for standardization. The standard approved by the American Standards Association in November, 1942, reduced the number of designs from many hundreds to 22 physical sizes. These cover the entire range used in both radio receivers and radio transmitters.

(2) H. P. Westman, "Mica-capacitator standard—First American war standard on radio," *Industrial Standardization and Commercial Standards Monthly*, vol. 13, pp. 297–299; December, 1942.
(3) Ray C. Ellis, "United States multiplies radio output," *Electronic Industries*, vol. 1, pp. 48–49, 106; November, 1942.
(4) "War standards for military radio," *Electronic Industries*, vol. 1, pp. 33–35; December, 1942.
(5) "Electronics—Secret weapon of war—Presager of a new scientific and industrial era," *Electronics*, vol. 15, facing p. 40; December, 1942.

* Decimal classification: R090. Original manuscript received by the Institute, February 1, 1943. This report is based on material from the 1942 Annual Review Committee of the Institute of Radio Engineers, as co-ordinated and edited by Laurens E. Whittmore, Keith Henney, and Frederick B. Llewellyn.

Radio Transmitters

General

The use of critical materials and vacuum tubes in standard and high-frequency broadcasting and in commercial television was limited to maintenance and repair of existing facilities. In general, new commercial developments in these fields have been retarded in order that manpower and materials could be concentrated in more urgently needed developmental activities directly connected with the prosecution of the war. Attention has been centered on extending the life of transmitter components—particularly transmitting tubes. All licensees of international broadcast stations in the United States entered into a closely co-ordinated plan of operation supervised and controlled by the Office of War Information and the Co-ordinator of Inter-American Affairs.

TABLE I

RADIO BROADCAST STATIONS FOR WHICH LICENSES AND CONSTRUCTION PERMITS ISSUED BY THE FEDERAL COMMUNICATIONS COMMISSION WERE OUTSTANDING ON DECEMBER 31, 1942

Class of Broadcast Station	Number of Licenses	Number of Construction Permits
Standard	910	7
Commercial High-Frequency (Frequency-Modulation)	11	36
Experimental High Frequency (Including 10 stations operating under "special authorization")	12	—
Commercial Television	3	5
Experimental Television	20	11
International	13	1
Facsimile	3	—
Noncommercial Educational	5	2

Standard-Band Broadcasting

The trend toward higher power in broadcast stations in the United States continued briefly until all new permits for such construction were discontinued and most outstanding permits for new facilities were canceled, to save critical war materials.

Experimental operation of a 1000-watt polyphase broadcast system was described. It was shown that performance comparable to that of other commonly used types of broadcast transmitting equipment could be obtained readily. The theoretically obtainable economies in tubes and equipment were realized. An improved design of audio-frequency phase-shifting network was completed which would limit the departure to 3 degrees from the desired 90 degrees over the range of 20 to 12,000 cycles per second.

(6) Paul Loyet, "Experimental polyphase broadcasting," PROC. I.R.E., vol. 30, pp. 213–222; May, 1942.

Two papers described calculating machines for tracing the field pattern of 2- or 3-element directional antenna arrays. In designing directional arrays many variables must be considered, such as spacing, phasing and current ratios, which make the calculations very laborious. The machines described eliminate the major part of the tedious work.

(7) F. Alton Everest and Wilson S. Pritchett, "Horizontal-polar-pattern tracer for directional broadcast antennas," PROC. I.R.E., vol. 30, pp. 227–232; May, 1942.

(8) William G. Hutton and R. Morris Pierce, "A mechanical calculator for directional antenna patterns," PROC. I.R.E., vol. 30, pp. 233–237; May, 1942.

Frequency-Modulation Broadcasting

Frequency-modulation broadcasting continued to expand, as new stations were completed and new receivers sold, until new construction was stopped by wartime restrictions on the use of critical materials. The industry standardized transmitter powers at $\frac{1}{4}$, 1, 3, 10, 25, 50, and 100 kilowatts at the instigation of the Federal Communications Commission. All sizes were in production except the 25- and 100-kilowatt transmitters.

International Broadcast Transmitters

A 100-kilowatt plate-modulated international broadcast transmitter was placed in operation at San Francisco for the transmission of broadcast programs to South America and the Orient.

Transmitters for Frequency-Modulation Broadcasting and Studio-to-Transmitter (S-T) Circuits

Several frequency-modulation broadcast transmitters for the range of 42 to 50 megacycles, manufactured in 1941, were placed in commercial operation in 1942. Popular output ratings were 1, 3, and 10 kilowatts. Also, several 25-watt frequency-modulation studio-to-transmitter link transmitters went on the air for the first time in the frequency range 330.4 to 343.6 megacycles, the application being the high-fidelity transmission of programs from studio to main transmitter.

(9) W. F. Goetter, "Frequency-modulation transmitter-receiver for studio-to-transmitter relay system," (abstract), PROC. I.R.E., vol. 30, pp. 251–252; May, 1942.

Frequency-Modulation Broadcast Antennas

The development of the circular or "doughnut" frequency-modulation broadcast antenna (frequency range 42 to 50 megacycles) was carried forward during the latter part of 1942 and a few units were placed in regular operation. This type of antenna has a circular field pattern, and the field is concentrated in the horizontal plane. This antenna may be readily adjusted to any frequency in the band without major physical change.

(10) M. W. Scheldorf, "Circular antenna," (abstract), PROC. I.R.E., vol. 30, p. 253; May, 1942.

Frequency-Modulation Transmitters and Receivers for Emergency Service

The use of frequency modulation for two-way emergency radio communication (30 to 42 megacycles) increased tremendously in 1942, since most of the applications are closely allied with the war effort. Standard output ratings are 30 watts for mobile service

and 60 or 250 watts for fixed and land stations. Special military designs have been built, incorporating features which gives the utmost in reliability under severe climatic conditions. Unattended relay stations have been placed in service in the band 116 to 119 megacycles with 10 watts output power. Such relay stations are needed when propagation conditions will not permit direct communication between the mobile station and headquarters.

(11) Herbert DuVal, Jr., "The tests that proved FM vital to communications," *Gen. Elec. Rev.*, vol. 22, pp. 5–7; February, 1942.

Communication with Automobiles

A most interesting application of both frequency modulation and amplitude modulation in the 33- to 38-megacycle and 116- to 119-megacycle bands was described in connection with the communications system of the Pennsylvania Turnpike which extends 161 miles between Pittsburgh and Harrisburg. The system provides two-way simplex communication between any of thirty patrol or official automobiles, seventeen ticket booths, ten utility buildings, six maintenance buildings and patrol headquarters, with connections to the Turnpike Commission headquarters and radio station in Harrisburg. Transmissions from any fixed point or mobile station are audible not only to the desired receiving terminal but simultaneously to all others in the system.

Six 116- to 119-megacycle amplitude-modulated automatically operated repeater stations carry traffic from end to end of the system. Mobile and branch units operate in the 33- to 38-megacycle band.

(12) "Pennsylvania Turnpike UHF traffic control system," *Electronics*, vol. 15, pp. 34–51; May, 1942.

Marine Service

A high-frequency marine radio transmitter of the unit type was introduced commercially during the year. This is a companion unit to the intermediate-frequency marine radio units which are now being installed on merchant ships by two major radio organizations. The new unit, like its intermediate-frequency counterpart, is a completely factory-assembled ship's radio station for the frequency bands covered, requiring connection only of power and antenna leads, thus saving thousands of hours of valuable shipyard labor.

Electronics

Cathode-Ray and Television Tubes

Military interest in cathode-ray tubes continued high during 1942. Of the matters which can be reported, mention should be made of the establishment of electrical and mechanical specifications by the Cathode-Ray Tube Committee of the Radio Manufacturers Association covering a group of preferred cathode-ray-tube types. The magnetic-deflection-tube types in this group employ a new electron gun with improved focus and have common structural and operating characteristics. The electrostatic-deflection types have increased spacing between high- and low-voltage leads, employ new base and socket designs, and have separate leads to all deflection plates for balanced deflection. After-acceleration is used to increase the light output and to improve the focus.

The development of the square-wave oscillograph and its application to the study of response characteristics of television apparatus represents an important contribution to work in this field, and provides a more accurate means of specifying the performance of television equipment.

A report on the relative sensitivities of television pickup tubes, photographic film, and the human eye gives an interesting comparison on a fundamental basis of the operating sensitivities of common methods of picture pickup. The relations between the operating sensitivities and the resolutions of the different methods are covered.

The use of the electron microscope has been extended during the past year and simpler portable models were demonstrated.

Important contributions were published in the field of electron optics.

(13) A. V. Bedford and G. L. Fredendall, "Analysis, synthesis, and evaluation of the transient response of television apparatus," Proc. I.R.E., vol. 30, pp. 440–457; October, 1942.
(14) R. D. Kell, A. V. Bedford, and H. N. Kozanowski, "A portable high-frequency square-wave oscillograph for television," Proc. I.R.E., vol. 30, pp. 45–8–464; October, 1942.
(15) V. J. Schaefer and D. Harker, "Surface replicas for use in the electron miscroscope," *Jour. Appl. Phys.*, vol. 13, pp. 427–433; July, 1942.
(16) R. F. Baker, E. G. Ramberg, and J. Hillier, "The photographic action of electrons in the range between 40 and 212 kv," *Jour. Appl. Phys.*, vol. 13, pp. 450–456; July, 1942.
(17) Henry Green, "The physics of pigments in dispersed systems," *Jour. Appl. Phys.*, vol. 13, pp. 611–622; October, 1942.
(18) Gilbert N. Plass, "Electrostatic electron lenses with a minimum of spherical aberration," *Jour. Appl. Phys.*; vol. 13, pp. 49–55; January, 1942.
(19) R. R. Law, "Factors governing performance of electron guns in television cathode-ray tubes," Proc. I.R.E., vol. 30, pp. 103–105; February, 1942.
(20) Karl Spangenberg and Lester M. Field, "Some simplified methods of determining the optical characteristics of electron lenses," Proc. I.R.E., vol. 30, pp. 138–144; March, 1942.
(21) Sidney Bertram, "Calculation of axially symmetric fields," *Jour. Appl. Phys.*, vol. 13, pp. 496–502; August, 1942.
(22) E. G. Ramberg, "Variation of the axial aberrations of electron lenses with lens strength," *Jour. Appl. Phys.*, vol. 13, pp. 582–594; September, 1942.

Gas-Filled Tubes

A contribution has been made as to the understanding of the phenomena of the arcback in mercury-vapor thyratrons.

(23) A. W. Hull and Frank R. Elder, "The phase of arcback," *Jour. Appl. Phys.*, vol. 13, pp. 171–178; March, 1942.

Small High-Vacuum Tubes

During the year 1942, a limited number of new receiving-tube types were introduced. Outside of the fact that some of these tubes were intended to operate at

higher than conventional receiver frequencies, and in some cases were intended for both receiving and transmitting applications, their designs were along conventional lines. In the United States a record number of small high-vacuum tubes was sold for the year with increased production on a reduced number of tube types. The total number of tubes manufactured during the year for commercial purposes was very substantially curtailed by the diversion of tube-manufacturing facilities to military requirements.

A vacuum device having a trigger property similar to that of a thyratron but obtaining this property by the use of the secondary emitting property of one of its elements has been described. This device differs in its characteristics from a thyratron in that its plate current can be stopped as well as started by the control grid. It will also function at much higher frequencies than a gas-filled device because it is not limited in its frequency by deionization time.

(24) A. M. Skellett, "The use of secondary electron emission to obtain trigger or relay action," *Jour. Appl. Phys.*, vol. 13, pp. 519–523; August, 1942.

Large High-Vacuum Tubes

In the field of large high-vacuum tubes, important developments in the materials substitution field have been made. Materials once considered indispensable in tube manufacture have been replaced quite successfully. Steel has replaced molybdenum and tungsten for many requirements. Zirconium-coated molybdenum and zirconium-coated graphite have found favor as anode materials. Similarly, stainless steel and nichrome and nickel-plated iron are being used for supports and shields. Ceramics have replaced natural-lava insulators.

Kovar has been utilized to replace copper for vacuum seals in high-power tubes operating at high frequencies. In the case of one large tube, the use of Kovar produced a temperature rise of only approximately 10 degrees Fahrenheit over the temperature of a copper-seal tube using the same quantity of cooling air.

It was reported that an oscillator circuit which minimizes the apparent effect of tube capacitance as a limitation to the maximum operating frequency of a tube has been developed.

(25) C. E. Haller, "The design and development of three new ultra-high-frequency transmitting tubes," Proc. I.R.E., vol. 30, pp. 20–26; January, 1942.

(26) I. E. Mouromtseff, "Water and forced-air cooling of vacuum tubes," Proc. I.R.E., vol. 30, pp. 190–205; April, 1942.

(27) E. L. Chaffee, "Characteristic curves of the triode," Proc. I.R.E., vol. 30, pp. 383–395; August, 1942.

(28) J. C. Frommer, "A graphical method to find the optimal operating conditions of triodes as class C telegraph transmitters," Proc. I.R.E., vol. 30, pp. 519–525; November, 1942.

(29) R. I. Sarbacher, "Graphical determination of power amplifier performance," *Electronics*, vol. 15, pp. 52–58, 158; December, 1942.

Television Broadcasting

The commercial television service that had been inaugurated in the United States in July, 1941, was continued into 1942, until May 25, when the Federal Communications Commission authorized a reduction in the mandatory program hours of commercial television stations from 15 to 4 hours per week. Up until this time, television-program technique had been developing rapidly with broadcasting of indoor and outdoor seasonal sports events, in addition to programs such as The Town Meeting of the Air from New York's Town Hall and the President's Birthday Ball from the Waldorf-Astoria. Live-talent studio programs were also regular features with dramatic stage shows and illustrated news programs.

In addition to the black-and-white programs carried out on regular National Television Standards Committee (NTSC) standards, one station also regularly broadcast color transmissions on the NTSC recommended color standards of 375 lines and 60 frames. Early in 1942 automatic color phasing was introduced in the color transmissions, so that color disks at the receivers could automatically lock into the proper phase.

Both the program hours and diversity of television broadcasts were sharply reduced after the curtailment in May, 1942. However, in the New York Area, NBC, CBS, and DuMont staggered their program schedules to give the televiewers an opportunity to see all of the program material still broadcast. A limited television-program service was also continued in Schenectady, Philadelphia, Los Angeles, and Chicago. Programs originating in New York have been rebroadcast regularly from Philadelphia and Schenectady via radio relay.

In the New York area it has been reported that there is considerably more interference from multipath effects in the 78- to 84-megacycle channel than has been observed in the 50- to 56-megacycle and the 60- to 66-megacycle channels.

Stereoscopic color television was demonstrated in England.

(30) B. J. Edwards, "Advantages and disadvantages of various types of focusing and deflection methods used in television," (abstract) *Radio News*, vol. 28, pp. 42–43; September, 1942.

(31) G. Wendt, "Image errors occurring in the deflection of a cathode-ray beam in two crossed deflecting fields" (abstract), *Wireless Eng.*, vol. 19, pp. 274–275; June, 1942.

(32) R. L. Campbell, R. E. Kessler, R. E. Rutherford, and K. U. Landsberg, "Mobile television equipment," Proc. I.R.E., vol. 30, pp. 1–7; January, 1942; *Jour. Soc. Mot. Pic. Eng.*, vol. 39, pp. 22–36; July, 1942.

(33) A. Rose, "Relative sensitivities of television pickup tubes, photographic film, and the human eye," Proc. I.R.E., vol. 30, pp. 293–300; June, 1942.

(34) O. S. Puckle, "Time bases," *Jour. I.E.E.* (London), vol. 89, pp. 100–119; June, 1942.

(35) E. C. Cherry, "Transmission characteristics of asymmetric-sideband communication networks," *Jour. I.E.E.* (London), vol. 89, pt. III, pp. 19–39 (discussion, pp. 39–42); March, 1942.

(36) R. B. Fuller and L. S. Rhodes, "Production of 16 mm. motion pictures for television projection," *Jour. Soc. Mot. Pic. Eng.*, vol. 39, pp. 195–201; September, 1942.

(37) G. L. Beers, "The focusing view-finder problem in television cameras," (abstract), Proc. I.R.E., vol. 30, p. 250, May, 1942; (abstract), *Electronics*, vol. 15, pp. 88–89; August, 1942; Proc. I.R.E., vol. 31, pp. 100–106; March, 1943.

(38) K. R. Wendt and G. L. Fredendall, "Automatic frequency and phase control of synchronization in television receivers," (abstract), Proc. I.R.E., vol. 30, p. 254; May, 1942; (abstract) *Electronics*, vol. 15, p. 89; August, 1942; Proc. I.R.E., vol. 31, pp. 7–15; January, 1943.

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(40) H. A. Breeding, "Mercury lighting for television studios," (abstract), Proc. I.R.E., vol. 30, p. 250; May, 1942; (abstract), *Electronics*, vol. 15, pp. 86–88; August, 1942; Proc. I.R.E., vol. 31, pp. 106–112; March, 1943.

(41) E. P. Anderson, "Audels New Radioman's Guide; Theory, Construction and Servicing; Television," 766 pages, Theodore Audel and Co., New York, N. Y., 1942.

(42) R. A. Monfort and F. J. Somers, "Measurement of the slope and duration of television synchronizing impulses," *RCA Rev.*, vol. 6, pp. 370–389; January, 1942.

(43) L. deForest, "Television Today and Tomorrow," 361 pages, Dial Press, New York, N. Y., 1942.

(44) P. C. Goldmark, J. N. Dyer, E. R. Piore, and J. M. Hollywood, "Color television—Part I," Proc. I.R.E., vol. 30, pp. 162–182; April, 1942; *Jour. Soc. Mot. Pic. Eng.*, vol. 38, pp. 311–352; April, 1942; *Jour. Appl. Phys.*, vol. 13, p. 666; November, 1942.

(45) J. L. Baird, "New television development; use of colour technique and stereoscopic relief," (abstract), *Electrician*, (London), vol. 127, p. 359; December 26, 1941.

(46) J. L. Baird, "Stereoscopic television in colour," (abstract), *Engineering*, (London), vol. 153, p. 35; January, 1942.

(47) J. L. Baird, "Three dimension color television," (abstract), *Electronics*, vol. 15, p. 76; May, 1942.

(48) R. Eichberg, "Photographing television images," *Amer. Photography*, vol. 36, pp. 8–12; April, 1942; pp. 16–18; May, 1942; pp. 22–24; June, 1942.

(49) R. Eichberg, "Adapting television receivers to the new standards," *Electronics*, vol. 14, p. 63; November, 1941.

(50) "Heat-treated television; electric heater protects antenna," *Gen. Elec. Rev.*, vol. 44, p. 696; December, 1941.

(51) W. L. Carlson, "Television reception with built-in antennas for horizontally and vertically polarized waves," *RCA Rev.*, vol. 6, pp. 443–454; April, 1942.

(52) T. Chew, "Analysis of feedback and prevention of oscillation in HF amplifiers," *Radio News*, vol. 27, pp. 14–15, 44–46; June, 1942.

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(54) H. L. Donley and D. W. Epstein, "Low frequency characteristics of the coupling circuits of single and multi-stage video amplifiers," *RCA Rev.*, vol. 6, pp. 416–433; April, 1942.

(55) M. A. Trainer, "Orthicon portable television equipment," Proc. I.R.E., vol. 30, pp. 15–19; January, 1942.

(56) J. B. Sherman, "A simple television demonstration system," Proc. I.R.E., vol. 30, pp. 8–15; January, 1942.

(57) F. Fischer and H. Thiemann, "Theoretical considerations on a new method of large-screen television projection," (abstract), *Wireless Eng.*, vol. 18, pp. 469–470; November, 1941.

(58) J. A. Rankin, "Receiver input connections for U-H-F-measurements," *RCA Rev.*, vol. 6, pp. 473–481; April, 1942.

(59) "Television light supplemented with fluorescent tubes," *Sci. Amer.*, vol. 166, pp. 195–196; April, 1942.

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(62) C. E. Lockhart, "Television waveforms," *Electronic Eng.*, vol. 15, pp. 19–26; June, 1942.

(63) "Television in colour and stereoscopic relief," *Jour. Telev. Soc.*, vol. 3, pp. 225–226; 1941–1942.

(64) M. E. Strieby and J. F. Wentz, "Television transmission over wire lines," *Jour. Telev. Soc.*, vol. 3, pp. 212–215; 1941–1942.

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(66) B. J. Edwards, "Design of television receiving apparatus," *Jour. Telev. Soc.*, vol. 3, pp. 204–211; 1941–1942.

(67) J. O. Edson, "A ten megacycle oscilloscope," *Bell. Lab. Rec.*, vol. 20, pp. 95–98; December, 1941.

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(69) R. Mautrer and F. Somers, "An experimental amateur television system," *Electronics*, vol. 15, pp. 68–71, 170–173; December, 1942.

(70) H. B. Fancher, "High-power television transmitter," (abstract), Proc. I.R.E., vol. 30, p. 251; May, 1942.

(71) J. E. Keister, "A television relay system," (abstract), Proc. I.R.E., vol. 30, p. 252; May, 1942.

Piezoelectricity

Workers in this field have been devoting practically all of their attention to developments and production for military purposes.

An interesting development in high-frequency crystal oscillators was described. By utilizing mechanical harmonics of crystals of practical size, oscillations of 197 megacycles, and harmonics as high as the twenty-third have been obtained with normal temperature-frequency characteristics. It was found that the Q of a crystal is independent of frequency but in general with increasing harmonic order. It was shown, in a discussion of oscillator circuits, that a capacitance-bridge oscillator circuit, utilizing the crystal in one arm, is best for high-frequency harmonic crystals because it annuls the large shunt capacitance of the crystal which prevents it from having the necessary positive reactance near the resonant frequency.

(72) W. P. Mason and I. E. Fair, "A new direct crystal-controlled oscillator for ultra-short-wave frequencies," Proc. I.R.E., vol. 30, pp. 464–472; October, 1942.

A treatment of the use of crystals as transducers is included in a book published during the year.

(73) W. P. Mason, "Electro-mechanical transducers and wave filters," 333 pages, D. Van Nostrand Company, Inc., New York, N. Y.; 1942.

Institute Affairs

New York Section

A New York Section of the Institute was formed during 1942, having its own officers as in the case of the other 26 Sections. Meetings of this Section for the presentation of technical papers now supersede the technical meetings of the Institute previously held in New York.

Standards Reports

The Institute Standards have been published in large-size individual pamphlets suitable for insertion in loose-leaf notebooks. These include the following:

Facsimile—Definitions of Terms (1942)

Radio Receivers—Definitions of Terms (Reprinted from 1938)

Radio Receivers—Methods of Testing Broadcast Radio Receivers (Reprinted from 1938. Spanish translation also available)

Radio Wave Propagation—Definitions of Terms (1942)

Radio Wave Propagation—Measuring Methods (1942)

Transmitters and Antennas—Definitions of Terms (Reprinted from 1938)

Transmitters and Antennas—Methods of Testing (Reprinted from 1938)

Electroacoustics—Definitions of Terms (Reprinted from 1938)

Acknowledgments

This summary of progress during 1942 covers, in general, the period up to the first of November, and is necessarily limited chiefly to certain developments in the United States which can appropriately be published under present circumstances. It is based on material prepared by members of the 1942 Annual Review Committee of the Institute of Radio Engineers. The final editing and co-ordinating was carried out by Laurens E. Whittemore, chairman; Keith Henney, and F. B. Llewellyn.

The other members of the Annual Review Committee for 1942 and the committees of the Institute of which they are chairmen are as follows:

E. G. Ports —Transmitters and Antennas
R. S. Burnap —Electronics
I. J. Kaar —Television
D. E. Noble —Frequency Modulation
J. H. Dellinger—Radio Wave Propagation
W. M. Angus —Radio Receivers
J. L. Callahan —Facsimile
W. G. Cady —Piezoelectric Crystals
G. G. Muller —Electroacoustics
H. A. Wheeler —Standards
O. H. Caldwell—Public Relations

The chairmen of the above committees wish to acknowledge the assistance given them by individual members of the committees.

Frequency-Modulation Distortion in Loudspeakers*

G. L. BEERS†, MEMBER, AND H. BELAR†, NONMEMBER, I.R.E.

***Summary*—As the frequency-response-range of a sound-reproducing system is extended, the necessity for minimizing all forms of distortion is correspondingly increased. The part which the loudspeaker can contribute to the over-all distortion of a reproducing system has been frequently considered. A type of loudspeaker distortion which has not received general consideration is described. This distortion is a result of the Doppler effect and produces frequency modulation in loudspeakers reproducing complex tones. Equations for this type of distortion are given. Measurements which confirm the calculated distortion in several loudspeakers are shown. The first Appendix giving the derivation of the equations is included.**

Introduction

Distortion in its several forms is undoubtedly the most serious problem with which the engineer has to deal in the design of a satisfactory high-fidelity sound-reproducing system. An early appreciation of the importance of minimizing distortion in such a system was obtained in development work on high-fidelity radio receivers during the period from 1930 to 1935. At that time an experimental radio receiver was developed which gave an over-all distortion from modulated radio-frequency input to audio-frequency output at the loudspeaker terminals of 2 to 3 per cent arithmetic sum of all harmonics. During the development of this receiver, consideration was given to the part which the loudspeaker might contribute to the over-all distortion in a sound-reproducing system. The possibility of a loudspeaker, reproducing a complex wave, creating frequency-modulation distortion by virtue of the Doppler effect was suggested. However, no facilities were available at that time to determine the extent of this type of distortion.

In the past few years frequency-modulation broadcasting and improved record reproducing systems have renewed the interest in high-fidelity sound reproduction. During a recent review of the distortion problem, a mathematical and experimental investigation of the possibility of frequency-modulation distortion in loudspeakers was conducted. It is the purpose of this paper to record the results of this investigation.

Theoretical Considerations

A loudspeaker cone mounted on a flat baffle may have to move through a relatively large distance at low frequencies to produce even a small acoustic output. If such a loudspeaker cone is simultaneously radiating both a low and a high frequency the source of high-frequency energy can be considered as moving back and forth in space and the high-frequency energy will therefore be frequency-modulated. This modulation in

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† RCA Manufacturing Company, Camden, New Jersey.

frequency is due to the Doppler effect and can be analyzed by assuming the speaker to be a constant-tone generator of the modulated frequency which is being displaced along the axis of the cone at the rate and amplitude of the modulating frequency. The resulting frequency-modulated wave can be represented by a "carrier" and an infinite number of sidebands. The fundamental or carrier is of constant amplitude and frequency. Its amplitude is less than that of the original wave and depends upon the degree of modulation. The sidebands are located symmetrically about the carrier and are spaced at intervals equal to the modulating frequency. The distortion introduced by these sidebands can be expressed by a distortion factor. Along the axis of a cone speaker this distortion factor can be stated as

$$\text{d.f.} = 0.033A_1f_2 \text{ per cent}$$

where

A_1 = amplitude of cone motion (each side of mean position) in inches at the modulating frequency.

f_2 = modulated frequency.

d.f. = distortion factor in per cent defined as the square root of the ratio of power in the sidebands to total power in the wave.

As shown in Appendix II, the amplitude of the cone motion required to radiate a given acoustic power varies inversely with the square of the frequency and inversely with the square of the diameter of the speaker cone.

The amplitude of the cone motion is likewise proportional to the square root of the audio-frequency power supplied to the loudspeaker. A 12-inch cone working with both sides in air requires a motion through a peak amplitude of approximately 1/16 inch (each side of its mean position) to radiate 1 acoustic watt at 100 cycles. Thus it can be said that a 12-inch cone radiating 1 acoustic watt at 100 cycles simultaneously with a 5000-cycle signal will distort the 5000-cycle signal by approximately 10 per cent due to frequency modulation. An 8-inch cone could only radiate 0.21 watt at 100 cycles for the same distortion at 5000 cycles and only 0.013 watt at 50 cycles for the same distortion limit.

The distortion factor, for any other low frequency f_1, cone size, and power radiated, is given by the following formula: assuming a flat baffle with the cone radiating on both sides.

$$\text{d.f.} = 2900\,\frac{f_2\sqrt{P_1}}{f_1^2d^2} \text{ per cent}$$

where f_2 = modulated frequency

f_1 = modulating frequency

P_1 = acoustic power output at f_1 in watts

d = cone diameter in inches

Comparisons with Similar Forms of Distortion

A distortion, similar to frequency modulation of complex tones in loudspeakers, occurs in the reproduction of sound from film, due to sprocket-hole modulation, and in disk reproduction due to tracking angle. In the case of film reproduction a distortion factor of 10 per cent would be had with an unfiltered sound gate reproducing a 400-cycle signal with a 1-mil displacement at sprocket-hole frequency. In the case of a disk reproducer, 10 per cent distortion would be had with a straight $7\frac{1}{4}$-inch tone arm reproducing a 5000-cycle note simultaneously with a 100-cycle note of 0.5-mil amplitude.

Distortion Measurements

In making distortion measurements on several loudspeakers the following procedure was followed. The speaker under test was set up in the sound-measuring room and fed with a variable frequency from the measuring channel. In series with this signal, right at the voice coil, was also introduced a low-frequency signal from another oscillator. The resultant complex sound was picked up by a ribbon microphone two feet from the speaker and on the axis of the cone. The output of the microphone was fed to a distortion-factor meter through a high-pass filter which eliminated the low-frequency signal. In the distortion-factor meter the modulated or distorted signal from the high-pass filter was balanced against a portion of the same frequency fed there directly from the variable oscillator. The phase and amplitude of the balancing signal were adjusted until a minimum reading was obtained. The residual reading represented the root-mean-square voltage of all forms of distortion, terms which may be simple harmonic distortion, cross modulation due to nonuniformity of flux density in the air gap, frequency-modulation distortion, or any other forms of distortion. To eliminate the effect of noise, readings of noise were taken and measurements of distortion made only over the range for which noise was not a factor.

The effect of harmonic distortion was isolated by making distortion measurements with a single-frequency input at the same level. Objections can be raised against using a single microphone in a single position for distortion measurements. Any one reading of distortion is subject to error due to the uneven distribution of sound in a room. However, in this case, a number of readings were taken at various frequencies which together should show the trend of the distortion characteristics. If allowance is made for noise and harmonic distortion, the remaining distortion of a complex sound would be made up of both cross modulation and frequency modulation. Since both produce sidebands of the same frequencies they cannot be separated easily for any one reading, but the amount due to each can be determined by measuring the distortion at various frequencies. The cross-modulation distortion should be proportional to the amplitudes

and should therefore be independent of frequency, whereas the frequency-modulation should be proportional to the frequency because the phase shift at shorter wavelengths is greater for the same mechanical displacement. Measurements were made of the mechanical displacement of the cone by means of a depth gauge which was adjusted to touch a small aluminum cone fastened to the voice coil for the purpose of this measurement.

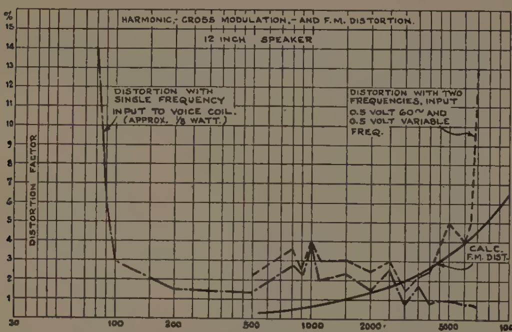

Fig. 1—Harmonic cross-modulation and frequency-modulation distortion of 12-inch speaker.

It should be noted that the power input to the loudspeakers which was used in making the measurements of frequency-modulation distortion was 0.5 watt or less. Since all but the smallest table model or portable receivers are capable of supplying power outputs in excess of this value it is apparent that the measurements are indicative of conditions which are actually met in practice. In the case of loudspeakers used in public-address systems and in theaters the power used is many times this value.

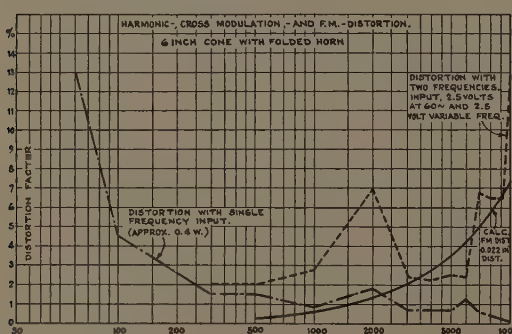

Fig. 2—Harmonic cross-modulation and frequency-modulation distortion of 6-inch cone with folded horn.

The calculated frequency-modulation distortion for the measured amplitude of cone displacement is shown on Fig. 1 for a conventional 12-inch cone speaker. It may be noted that at 5000 cycles and above, frequency-modulation distortion begins to have an effect; the calculated amount being 4 per cent at 6000 cycles. The measured amount follows the curve except that at 7000 cycles the measured distortion is much higher than that calculated. This may be explained by the fact that for the calculated frequency-modulation distortion, uniform response and a dead auditorium have been assumed. Frequency modulation of an audio wave becomes more noticeable in a live room. In such a room the pattern of standing waves is continuously shifted. Thus, small changes in phase can produce large changes in amplitude, thereby converting frequency modulation into amplitude modulation.

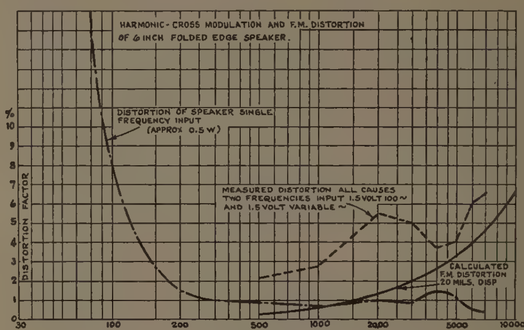

Fig. 3—Harmonic cross-modulation and frequency-modulation distortion of 6-inch folded-edge speaker.

Similar results were obtained with a 6-inch cone in a folded horn as shown in Fig. 2. Compared with the 12-inch speaker this horn and cone combination has less distortion below 100 cycles probably due to the loading of the horn. Otherwise the same rise in over-all distortion is noted at the high-frequency end which may be accounted for by frequency modulation. However, at 2000 cycles the particular unit tested had considerable cross-modulation distortion.

The characteristics of a 6-inch cone speaker of the folded-edge type are shown in Fig. 3. The distortion with a single tone is low at high frequencies but is

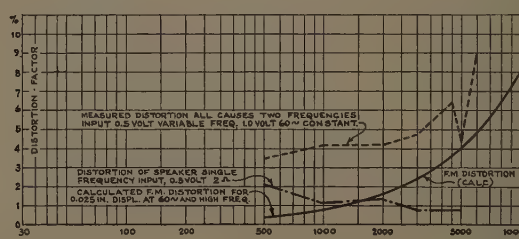

Fig. 4—Harmonic cross-modulation and frequency-modulation distortion of 12-inch cone, permanent-magnet field speaker.

always high when two tones are reproduced simultaneously. There is a trend toward greater distortion at higher frequencies in accordance with the calculated frequency-modulation distortion. The modulating frequency in this case was 100 cycles. At 60 cycles the distortion was too great to permit useful measurements.

Another speaker which also had considerable cross-modulation distortion in the midrange was a 12-inch permanent-magnet speaker. The characteristics of this

speaker are shown in Fig. 4. The increase in the over-all distortion with frequency, however, is still noticeable.

In order to test the distortion of the measuring channel, distortion measurements were made with the distortion meter connected directly across the voice coil of the speaker. This distortion always measured less than 1 per cent. However, this did not preclude the possibility of distortion in the microphone.

The latter is usually assumed to have a very low distortion. This was further verified by tests made on a combination speaker which employed separate high- and low-frequency units and which should have very low cross-modulation and frequency-modulation distortion. Measurements, as shown in Fig. 5, confirm this conclusion. The mid-frequency and high-frequency distortion with complex tones, with the exception of one point, is less than 2 per cent. The low-frequency distortion with a single tone showed a peak at 90 cycles but below that frequency the distortion is less than that of other speakers tested.

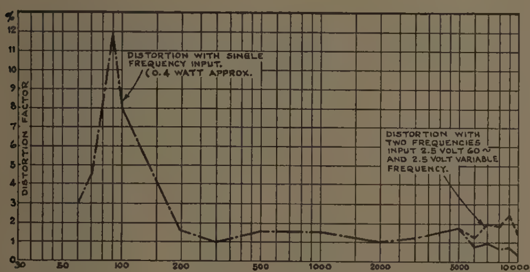

Fig. 5—Harmonic and cross-modulation distortion, separate low- and high-frequency speakers.

The frequency-modulation distortion should be independent of the amplitude of the modulated frequency but be directly proportional to the amplitude of the modulating frequency. Thus if 5000 cycles and 60 cycles are reproduced together the per cent distortion of the 5000-cycle wave due to frequency modulation should be independent of the amplitude at 5000 cycles but should be directly proportional to the amplitude of motion at 60 cycles. This is borne out by measurements shown in Figs. 6 and 7. Fig. 6 shows the measured distortion with the 60-cycle input varied. The increase in distortion with 60-cycle input may be noted. Fig. 7 shows the distortion for constant 60-cycle input but with the amplitude of the 5000-cycle signal varied. The per cent distortion should remain constant according to calculations, and the measurements confirmed this conclusion.

Listening Tests

A question which naturally arises is; how audible is this form of distortion and how much of it can be allowed for various applications?

Listening tests were made as follows: The signal and the modulating tone were applied to a speaker in the sound-measurement room and a microphone placed before it. The output of the microphone was fed to another speaker in another room through an amplifier and filter which eliminated the low-frequency modulating tone. Thus the observer could not hear the low-frequency tone itself but only its effect upon the higher frequency. As expected and explained before, the audibility of distortion was considerably increased if the microphone were used to pick up a greater amount of reflected sound. The audibility varied, but in general

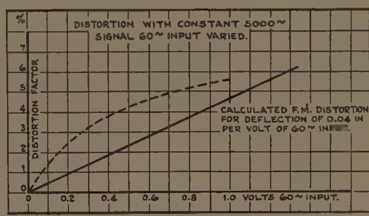

Fig. 6—Distortion of a complex tone (60 cycles and 5000 cycles) by a 12-inch speaker.

it can be said that a distortion of about 2 to 3 per cent becomes noticeable as a change in quality. The sensation produced is that of a very familiar form of distortion which is still hard to describe. The sound is just not clean.

Considerable difficulty arises in endeavoring to evaluate the objectionability of this distortion through listening tests involving the reproduction of speech and music. The most obvious method of obtaining the desired information is by a comparison between a speaker which is producing the distortion and one which is not. For such a comparison to be of real value it is necessary not only that the performance characteristics of the two speakers be substantially identical (except with

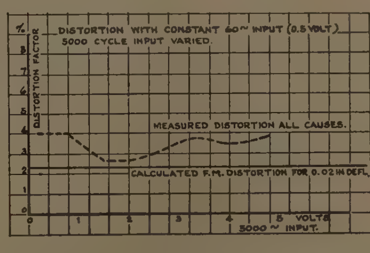

Fig. 7—Distortion by a 12-inch speaker with constant 60-cycle input and 5000-cycle input varied.

respect to the frequency-modulation distortion) but also that the distortion from all other sources be kept as low as possible. Although preliminary listening tests have indicated the presence of this distortion in the reproduction of music, the preference given other work, which at the present time is more essential, has prevented setting up the necessary facilities to conduct conclusive tests. It is hoped that these tests can be conducted in the near future.

Means of Reducing Frequency-Modulation Distortion of Complex Tones

Some of the means of reducing the distortion are as follows:

(1) Reduce the amplitude of cone traverse by loading the speaker with a horn.
(2) Increase the cone diameter and thus reduce amplitude of motion.
(3) Limit the power input at low frequencies.
(4) Use separate speakers for the low and high frequencies.

The following example illustrates the reduction in distortion which can be obtained by the use of separate low- and high-frequency speakers. The distortion of an 8-inch speaker in a flat baffle radiating 0.1 watt at 50 cycles simultaneously with a 10,000-cycle signal would be

$$\text{d.f.} = 2900\,\frac{f_2\sqrt{P}}{f_1^2 d^2} = 2900\,\frac{10{,}000\sqrt{0.1}}{50^2\times 8^2}$$
$$= 57 \text{ per cent approximately}$$

but if two 8-inch speakers were used divided at

$$f = \sqrt[3]{f_1^2 f_2} = \sqrt[3]{50^2\times 10{,}000} = 292 \text{ cycles}$$

the distortion under the assumed conditions would become

$$\text{d.f.} = 2900\,\frac{10{,}000\sqrt{0.1}}{292^2\times 8^2} = 2900\,\frac{292\sqrt{0.1}}{50^2\times 8^2}$$
$$= 1.7 \text{ per cent}$$

or a reduction of 34:1.

It may be noted that consideration has been given only for the distortion at a point along the axis of the cone. At right angles to the cone, frequency modulation of the type analyzed does not exist. In a normal listening room, however, sound radiated along the axis will reach all points of the room because of reflections. A more accurate analysis would integrate the output of fundamental and sidebands for any position in the room. It was not thought necessary, however, to do this to illustrate the effect.

Conclusions

Both a mathematical analysis and measurements indicate the possibility of frequency-modulation distortion in loudspeakers when reproducing a complex sound wave. Since this distortion increases with frequency its effects are most pronounced in high-fidelity reproducing systems. It is fortunate that several relatively simple means can be employed to minimize or avoid this type of distortion. It is difficult to determine its seriousness on the basis of listening tests because it cannot be isolated readily from other types of distortion. Although it is probable that this form of distortion is usually masked by distortion from other sources, nevertheless, it is a factor which should be considered if the maximum in faithful sound reproduction is desired.

Appendix I

Derivation of Equation for the Distortion Factor

The equation for the Doppler effect is[1]

$$f_0 = \frac{v - v_0}{v - v_s} f_s \tag{1}$$

where v = velocity of sound in medium
v_0 = velocity of observer
v_s = velocity of source
f_s = source frequency
f_0 = observed frequency.

Let f_1 = frequency of motion of source
f_2 = source frequency.

Assume the motion of the source to be

$$S = A_1 \sin 2\pi f_1 t. \tag{2}$$

The velocity of the source is

$$\dot{S} = 2\pi f_1 A_1 \cos 2\pi f_1 t.$$

Since the observer is at rest, $v_0 = 0$ and

$$f_0 = \frac{f_2}{1 - \dfrac{2\pi f_1 A_1}{v}\cos 2\pi f_1 t}$$
$$\approx f_2\left(1 + \frac{2\pi A_1}{\lambda_1}\cos 2\pi f_1 t\right) \tag{3}$$

where $v = f_1\lambda_1$ (λ_1 being the wavelength of the wave of frequency f_1) and $2\pi f_1 A_1 \ll v$.

In a uniform sine wave, such as $\sin \omega t$, the frequency is $1/2\pi$ times the rate of change of the argument of the sine function, or

$$f = \frac{1}{2\pi}\frac{d}{dt}(\omega t) = \frac{\omega}{2\pi}. \tag{4}$$

If the sine function is not uniform, it is customary to define its instantaneous *generalized frequency* in an analogous manner, as $1/2\pi$ times the rate of change of the argument of the sine function.[2,3]

$$f_0 = \frac{1}{2\pi}\frac{d}{dt}(\text{argument of sine function})$$

so

$$\text{argument of sine function} = 2\pi\int f_0\,dt$$
$$= 2\pi f_2\int\left\{1 + \frac{2\pi A_1}{\lambda_1}\cos 2\pi f_1 t\right\}dt$$
$$= 2\pi f_2 t + \frac{2\pi A_1}{\lambda_2}\sin 2\pi f_1 t \tag{5}$$

since $f_1\lambda_1 = f_2\lambda_2$.

[1] H. F. Olson, "Elements ot Accoustical Engineering," D. Van Nostrand Company, New York, N. Y., 1940, p. 16.
[2] Helmholtz, "Die Lehre von den Tonempfindungen," Braunschweig, 1913, pp. 649–650.
[3] John R. Carson, "Notes on the theory of modulation," Proc. I.R.E., vol. 10, p. 63; February, 1922.

The equation of a sound wave along the axis of a cone speaker, for a single tone, can be written

$$P = KA \sin\left(2\pi ft - \frac{2\pi X}{\lambda}\right) \tag{6}$$

where P = pressure

K = a constant depending upon the speaker design

A = amplitude of cone traverse

f = frequency

X = distance from speaker

λ = wavelength of the sound.

According to the Doppler effect, the movement of the source causes the frequency of the sound at X to vary. Equation (5) shows that the equation of the sound wave, when the motion of the source is considered, is

$$P = KA_2 \sin\left\{2\pi f_2 t + \frac{2\pi A_1}{\lambda_2}\sin 2\pi f_1 t - \frac{2\pi X}{\lambda_2}\right\} \tag{7}$$

where $2\pi X/\lambda_2$ is a constant phase-shift term which can be eliminated by choosing $X = n\lambda_2$.

The resultant equation is well known, and can be expressed in terms of its component frequencies by expansion in a series of Bessel functions as follows:[4]

$$\begin{aligned} P = KA_2\Big\{ & J_0\left(\frac{2\pi A_1}{\lambda_2}\right)\sin 2\pi f_2 t \\ & + J_1\left(\frac{2\pi A_1}{\lambda_2}\right)[\sin 2\pi(f_2+f_1)t - \sin 2\pi(f_2-f_1)t] \\ & + J_2\left(\frac{2\pi A_1}{\lambda_2}\right)[\sin 2\pi(f_2+2f_1)t + \sin 2\pi(f_2-2f_1)t] \\ & + J_3\left(\frac{2\pi A_1}{\lambda_2}\right)[\sin 2\pi(f_2+3f_1)t - \sin 2\pi(f_2-3f_1)t] \\ & + J_4 \cdots \Big\}. \end{aligned} \tag{8}$$

This means that the frequency-modulated wave consists of a fundamental or carrier which is constant and of frequency f_2 but with the amplitude reduced by the factor $J_0(2\pi A_1/\lambda_2)$.

Sidebands have been generated which are displaced from the carrier frequency by the amount of the frequency of motion of the source, and also by all the harmonics of this frequency as well. The amplitudes of the sideband are proportional to Bessel functions having the same order as the sideband.

The total power of the harmonic sidebands is proportional to

$$I_{\text{sideband}} = 2\sum_{n=1}^{n=\infty}\left[KA_2 J_n\left(\frac{2\pi A_1}{\lambda_2}\right)\right]^2. \tag{9}$$

[4] Balth. van der Pol, "Frequency modulation," Proc. I.R.E., vol. 18, p. 1199; July, 1930.

Since the total power of the wave is not changed by the frequency modulation,

$$I_{\text{total}} = (KA_2)^2\left[J_0\left(\frac{2\pi A_1}{\lambda_2}\right)\right]^2 + I_{\text{sideband}} = (KA_2)^2$$

so

$$I_{\text{sideband}} = I_{\text{total}}\left\{1 - \left[J_0\left(\frac{2\pi A_1}{\lambda_2}\right)\right]^2\right\}.$$

The distortion factor then becomes

$$\text{d.f.} = \sqrt{1 - \left[J_0\left(\frac{2\pi A_1}{\lambda_2}\right)\right]^2} \tag{10}$$

where the distortion factor has been defined as the ratio of the square root of the sideband power to the square root of the total power.

Since

$$J_0(X) = 1 - \frac{X^2}{2^2} + \frac{X^4}{2^2\cdot 4^2} - \frac{X^6}{2^2\cdot 4^2\cdot 6^2} + \cdots$$

$$1 - [J_0(X)]^2 = \frac{X^2}{2} - \frac{3X^4}{32} + \cdots$$

and the distortion factor is approximately

$$\text{d.f.} = \sqrt{1 - \left[J_0\left(\frac{2\pi A_1}{\lambda_2}\right)\right]^2} \approx \sqrt{\frac{1}{2}\left(\frac{2\pi A_1}{\lambda_2}\right)^2}$$

$$= 0.707\left(\frac{2\pi A_1}{\lambda_2}\right).$$

This approximation is very good for values of distortion up to 25 per cent. Since $\lambda_2 f_2 = C$ where $C = 1130$ feet per second, the expression becomes,

$$\text{d.f.} = 0.707(2\pi)\frac{f_2 A_1}{1130(12)}(100) \text{ per cent}$$

$$= 0.033 f_2 A_1 \text{ per cent.} \tag{11}$$

As shown in Appendix II, the amplitude of cone motion of a loudspeaker for a given power radiated is inversely proportional to the square of the frequency and the square of the cone diameter, and likewise is directly proportional to the square root of the audio-frequency power supplied.

$$A = K\frac{\sqrt{P}}{f^2 d^2}.$$

For a 12-inch cone working with both sides in air, and radiating 1 acoustic watt at 100 cycles, the amplitude of cone motion is approximately[5] 1/16 inch. This gives the equation for the distortion factor as

$$\text{d.f.} = 2900\frac{f_2\sqrt{P_1}}{f_1^2 d^2} \tag{12}$$

where d is in inches and P_1 is the acoustic power output in watts at f_1 cycles per second.

[5] Harold Pender and Knox McIlwain, "Electrical Engineers' Handbook," vol. 5, third edition, John Wiley and Sons, New York, N. Y., 1936, pp. 6–8.

Appendix II

Equation for the Amplitude of Motion of a Loudspeaker Cone

The mechanical impedance of the air load upon one side of a vibrating piston set in an infinite baffle is[6,7]

$$Z_M = \pi R^2 \rho c \left\{1 - \frac{J_1(2KR)}{KR}\right\} + j\frac{\pi\omega\rho}{2K^3}K_1(2KR)$$

where R = radius of piston
ρ = density of medium
c = velocity of sound in medium
$K = 2\pi/\lambda$
λ = wavelength of sound
$\omega = 2\pi f$
f = frequency, in cycles per second
$J_1(2KR)$ = Bessel's function of first kind
$K_1(2KR)$ = modified Bessel's function.

By the series expansion for Bessel's function of the first kind,

$$1 - \frac{J_1(2KR)}{KR} = \frac{K^2R^2}{2} - \frac{K^4R^4}{2^2(3)} + \frac{K^6R^6}{2^2(3^2)(4)} \cdots$$

so the mechanical resistance is given approximately by

$$r_{MA} = \pi R^2 \rho c \left\{\frac{K^2R^2}{2} - \frac{K^4R^4}{2^2(3)} + \frac{K^6R^6}{2^2(3^2)(4)}\right\}.$$

[6] Lord Rayleigh, "Theory of Sound," vol. 2. The Macmillan Company, New York, N. Y., 1926, p. 164.
[7] See page 80 of footnote reference 1.

The power output is given by

$$P = r_{MA}\dot{x}^2$$

where $\dot{x}$ = root-mean-square velocity of the piston.

If the piston motion is given by

$$x = A \sin \omega t$$

the velocity is

$$\dot{x} = A\omega \cos \omega t.$$

The power output is therefore

$$P = \pi R^2 \rho c A^2 \omega^2 \left\{\frac{K^2R^2}{2} - \frac{K^4R^4}{2^2(3)} + \frac{K^6R^6}{2^2(3^2)(4)}\right\} \cos^2 \omega t.$$

For low audio frequencies when $\lambda \gg R$, the average power output is given by

$$P = \frac{\pi R^2 \rho c A^2 \omega^2 K^2 R^2}{4}$$

since the first term of the series is much larger than succeeding ones and the average value of $\cos^2 \omega t$ is one half.

Solving for A, the cone amplitude

$$A = K\frac{\sqrt{P}}{f^2 d^2}$$

where K is a constant, f is the frequency in cycles per second, d is the cone diameter, and P is the radiated power.

This formula is not accurate at high audio frequencies since interference of the radiation from various parts of the disk occurs in space, causing a departure from spherical wave propagation.

Some Recent Developments in Record Reproducing Systems*

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Summary—Several factors of importance in obtaining satisfactory reproduction of sound from lateral-cut phonograph records are considered. An experimental record reproducing system employing the principles of frequency modulation is described and data are supplied on the measured and calculated performance characteristics of the system. Curves are included showing the vertical force required for satisfactory tracking with the experimental frequency-modulation pickup as compared with other pickups of conventional design.

Introduction

The remarkable increase in the sale of phonograph records during the past few years is a definite indication of the returning public interest in records as a medium of home entertainment. If the present interest in phonograph records is to be maintained, and on a permanent basis, it is essential that the reproduction of sound from records be at least comparable to or preferably better than that which can be obtained from radio broadcasting.

During the past few years, an investigation was conducted to determine the prospects of improving materially the over-all performance of record reproducing systems. One phase of the investigation was directed toward the possibility of reproducing frequencies up to 10,000 or 12,000 cycles from standard shellac records without the introduction of objectionable surface noise. In the course of this investigation, the possibilities of producing a frequency-modulated signal by means of a special pickup and associated circuits was investigated. This type of reproducing system appeared to lend itself to a realization of certain requirements

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which were considered essential to a satisfactory record reproducing system. It is the purpose of this paper to discuss some of these requirements, to describe the frequency-modulation record-reproducing system and indicate the improvement in performance which can be expected from such a system.

A. Pickup Requirements for Satisfactory Reproduction of Lateral-Cut Records

There are many factors which must be considered in the design of a pickup to reproduce lateral-cut records. The following factors not only determine the quality of reproduction which will be obtained but also have a direct bearing on the life of the record and stylus.

1. Vertical Force Required for Satisfactory Tracking

The vertical force required for satisfactory tracking should be low enough to prevent excessive record wear and minimize record surface noise. Numerous tests have been made which indicated that for lacquer records the vertical force should not exceed 20 grams. A maximum value of 30 grams is considered satisfactory for shellac records.

2. Mechanical Impedance

The vertical and lateral mechanical impedances presented by the pickup at the stylus should be as low as possible since the work which is performed by the record is a function of these impedances. Low mechanical impedance is likewise desirable to minimize the mechanical noise or chatter radiated directly from the pickup and record.

3. Free Resonance of Pickup

Experimental evidence indicates that it is desirable to keep the free resonance of the pickup at as high a frequency as possible to minimize the effect of ticks and other record-groove irregularities.

4. Relationship between Stylus Displacement and Audio Output

If distortion is to be minimized it is essential that the pickup and associated circuits provide a linear relationship between audio voltage output and stylus displacement. The necessity for minimizing distortion increases as the frequency range of a sound-reproducing system is extended.

5. Frequency-Response Characteristic

A phonograph pickup suitable for a high-fidelity system should provide a frequency response throughout the useful audio-frequency range which is proportional to either the amplitude or the velocity of the modulation in the record groove.

6. Sensitivity

The sensitivity of the pickup should be such that the amplification required between the pickup and loudspeaker is not so great as to introduce serious microphonic difficulties.

B. Frequency-Modulation Record Reproducing System

A review of the foregoing requirements led to a consideration of several types of record reproducing systems. The possibility of producing a frequency-modulated radio signal by means of a special pickup was investigated. It was found that with a simple pickup a frequency-modulated signal could be produced which had sufficient frequency deviation to provide a relatively high audio-frequency voltage output, when applied to a frequency discriminator and rectifier combination. Either the inductive or the capacitive reactance of a resonant circuit can be varied to produce a desired frequency shift. From the standpoint of a phonograph pickup the control of frequency through a variation in capacitance seemed to offer the greater advantage.

1. Frequency-Modulation Pickup

Fig. 1 shows in outline form the general construction of an experimental frequency-modulation pickup. A metal frame or mounting block is provided as a sup-

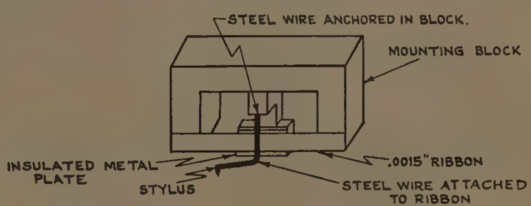

Fig. 1—General construction of frequency-modulation pickup.

port for an insulated plate which is the high-potential side of the pickup. To this mounting block is also attached a thin metal ribbon. The ribbon, which is mounted in a plane parallel to the insulated plate and spaced from it by a small air gap, is placed under tension in order to increase the natural resonance frequency of the system. The stylus supporting wire is anchored to the mounting block at its upper end. It is attached to the ribbon at approximately the mid-point of its length and its free end is bent in a plane essentially parallel to the record groove. The sapphire which is used as a stylus is attached to the end of the wire. The portion of the wire between the ribbon and the sapphire provides sufficient vertical compliance to minimize mechanical noise and to reduce distortion due to pinch effect. From the figure it is apparent that displacement of the stylus laterally results in a change in the position of the ribbon with respect to the fixed plate and thus produces a change in capacitance. The

over-all length of the mounting block shown in Fig. 1 is approximately $\frac{1}{2}$ inch. The normal spacing between the fixed plate and the ribbon is approximately 0.004 inch.

From a purely theoretical standpoint it is essential that in a frequency-modulation pickup the change in capacitance with displacement of the stylus be such as to produce a linear relationship between frequency change and motion of the stylus. In other words, the variable capacitor formed by the elements of the pickup should, in radio terminology, be of the straight-line frequency type. From a practical standpoint the distortion introduced by a pickup constructed along the lines indicated in Fig. 1, when used with circuits to be described later, is substantially negligible.

2. *Circuit Considerations*

The essential circuit considerations which are in volved in the design of a frequency-modulation record reproducing system may be stated as follows:

1. The carrier frequency to be employed.
2. A suitable oscillator circuit for use with the pickup.
3. The type of frequency-discriminator—rectifier combination to employ.

A study of the question of the operating frequency to use in a frequency-modulation phonograph system leads to the conclusion that carrier frequencies as low as those used in the intermediate-frequency amplifiers of radio receivers and as high as those employed for frequency-modulation broadcasting will give satisfactory results. If the phonograph is to be used in combination with a radio receiver there may be some advantage in using a carrier frequency which permits the use of one or more of the intermediate-frequency am-

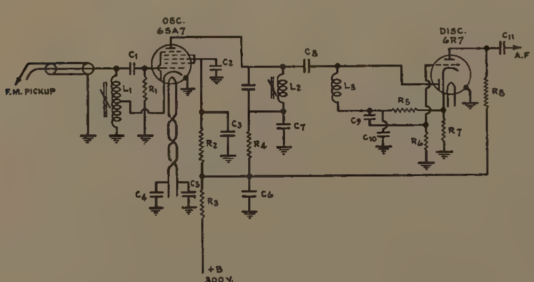

Fig. 2—Schematic diagram of oscillator-discriminator.

plifier circuits as a frequency-discriminating network for converting the frequency-modulated signal into amplitude modulation prior to detection. However, if the phonograph is designed as a separate device it may be desirable to use a frequency in the neighborhood of 30 megacycles particularly in case some frequency in this region is assigned by the Federal Communications Commission for diathermy machines. If a carrier frequency is used in a band thus allocated by the F.C.C. no special shielding would be required to prevent interference with other radio services. The signal level provided at the discriminator by the frequency-modulated oscillator can readily be made quite high, so there is no likelihood of diathermy machines or other electrical equipment causing interference with the phonograph.

Since the oscillator and frequency-discriminator-rectifier circuits are to a considerable extent interdependent, they will be discussed together. Fig. 2 is a schematic diagram of circuits which have given very satisfactory results. The circuit problem in connection with the oscillator is to provide an arrangement which will have sufficient frequency stability from the standpoint of line-voltage variations, temperature changes, etc., and at the same time enable the pickup capacitance variations to produce the desired frequency change.

From the standpoint of obtaining the maximum frequency change for a given variation in capacitance at the pickup, it is desirable that the pickup be connected directly across the oscillator tuned circuit. This can, of course, be accomplished by mounting the oscillator tube and associated circuit elements at the pickup end of the tone arm.

This arrangement has not been found to be particularly desirable because the tone arm is made unduly large and the heat from the oscillator tube causes the end of the tone arm, which is handled by the user, to become uncomfortably hot. The same result, however, can be accomplished by mounting the oscillator tube on the main instrument chassis and connecting it to the pickup through a resonant transmission line, which is used as the oscillator tuned circuit. It has been found that by connecting the pickup previously described through a relatively low capacitance line to a conventional oscillator circuit, as shown in the diagram, a sufficient frequency shift is obtained to give the desired audio-frequency output. In this case the transmission line is treated as a lumped capacitance. The line is included as an integral part of the tone arm.

It will be noted that the oscillator tube employed is of the 6SA7 type. This tube permits the use of electronic coupling between the oscillator and discriminator circuits. The oscillator frequency is adjustable by means of an iron core which is associated with the inductance L_1 shown in the diagram.

A simple resonant circuit is utilized as the means for converting the oscillator-frequency variations into changes in the amplitude of the signal applied to the diode portion of the 6R7 tube. A powdered-iron core associated with inductance L_2 is used to tune this circuit so that the mean oscillator frequency falls at approximately the 70 per cent response point on one side of the selectivity characteristic. The rectification of the radio-frequency signal by the diode develops an audio-frequency potential across the resistor R_5. This audio-frequency potential is then amplified by the triode

section of the 6R7. The output voltage which appears across resistor R_8 in the plate circuit of the 6R7 is applied to a suitable audio-frequency amplifier and loudspeaker.

The audio-frequency output of the circuit shown in Fig. 2 is a function of

1. The oscillator voltage applied to the discriminator,
2. The frequency variations in this voltage which are produced by the pickup,
3. The slope of the discriminator network,
4. The audio voltage gain obtained from the 6R7.

An experimental pickup employed in the circuit shown in Fig. 2 has given a root-mean-square potential of 6 to 8 volts across resistor R_8 when reproducing a 400-cycle record cut at a groove amplitude of 0.001 inch.

C. Performance Characteristics

In the course of development of the frequency-modulation pickup system, equations were derived for use in calculating the performance characteristics. Derivations of these equations will be found in the Appendixes. Through the use of these equations the following characteristics were calculated:

(a) Lateral mechanical impedance.
(b) Lateral force acting upon stylus.
(c) Response characteristic of pickup and tone arm.
(d) Tracking weight required to overcome vertical force due to lateral velocity.
(e) Tracking weights and relative outputs to be obtained with different radius styli.

For purposes of comparison measurements were made on an experimental pickup to determine the last three of these characteristics.

1. *Calculated Characteristics*

(a) Lateral Mechanical Impedance

If sine-wave motion of the stylus on the record (perfect tracking) is assumed and the tone arm is of rigid construction then the equivalent electrical diagram shown in Fig. 3 may be set up. Since at high frequencies

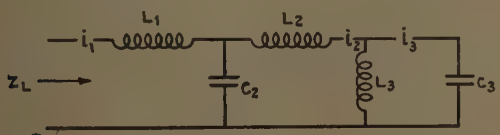

Fig. 3—Equivalent electrical diagram of frequency-modulation pickup and tone arm.

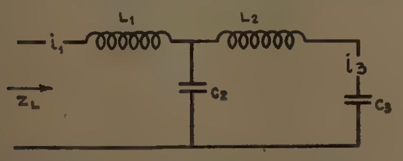

Fig. 4—Equivalent electrical diagram of frequency-modulation pickup for high-frequency calculations

mass L_3 (tone arm) is thousands of times higher than either the stylus or its supporting wire the equivalent diagram shown in Fig. 4 can be used. In this case the lateral impedance Z_{L_1} becomes

$$Z_{L_1} = \left(\frac{j}{\omega}\right)\frac{\omega^4 L_1 L_2 C_2 C_3 - \omega^2(L_1 C_3 + L_1 C_2 + L_2 C_3) + 1}{\omega^2 L_2 C_2 C_3 - (C_2 + C_3)}.$$

Analyzing the above equation, it will be found that the lateral impedance Z_{L_1} becomes infinite at $f=0$, $f=\infty$, and when $f=f_c$, where f_c is the frequency at which the pickup moving system resonates when the stylus is locked in the groove and may be expressed as

$$f_c = \frac{1}{2\pi\sqrt{L_2\dfrac{C_2C_3}{C_2+C_3}}}.$$

At low frequencies, we may again simplify by disregarding the stylus mass L_1 and the pickup moving system mass L_2. Fig. 5 shows the equivalent diagram for low frequencies.

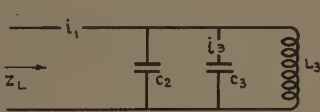

Fig. 5—Equivalent electrical diagram of frequency-modulation pickup and tone arm for low-frequency calculations.

In this case the lateral impedance Z_{L_2} becomes

$$Z_{L_2} = \frac{j\omega L_3}{1 - \omega^2 L_3(C_2 + C_3)}.$$

Analyzing this equation it is found that Z_{L_2} becomes infinite at f_s, the tone-arm swinging resonance where

$$f_s = \frac{1}{2\pi\sqrt{L_3(C_2+C_3)}}.$$

Fig. 6 shows the calculated impedance of the pickup and tone-arm system based on a pickup locked resonance of 15 kilocycles and a tone-arm swinging resonance of 20 cycles.

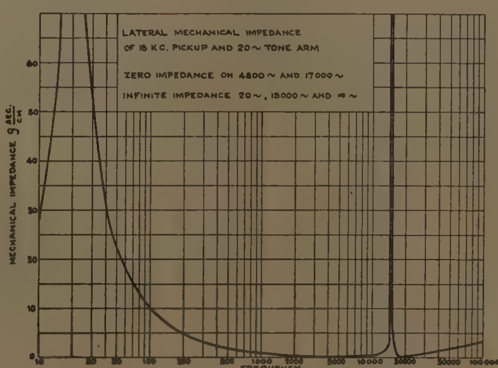

Fig. 6—Calculated lateral mechanical impedance of frequency-modulation pickup.

The three points of zero impedance will be noted at $f=0$, $f=4800$ cycles, and $f=17{,}000$ cycles. Two points of infinite impedance occur as shown at $f=20$ cycles and $f=15{,}000$ cycles.

(b) Lateral Force Acting upon Stylus

Using the calculated values of the effective mechanical impedance of the pickup and tone-arm system, it is possible to calculate the lateral force acting upon the

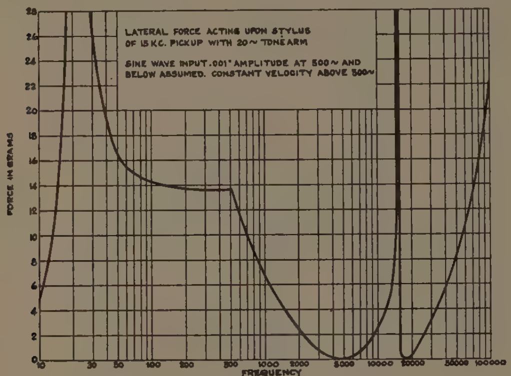

Fig. 7—Calculated lateral force acting upon the stylus of frequency-modulation pickup.

stylus due to record-groove modulation. For these calculations a groove modulation of 0.001 inch amplitude up to 500 cycles and a constant groove velocity above that frequency are assumed. Fig. 7 shows the curve obtained on the basis of these assumptions. It will be noted that the lateral force exerted on the stylus reaches a theoretically infinite value at 20 cycles and at 15,000 cycles, the resonant frequencies of the tone arm and pickup.

(c) Response Characteristic of Pickup and Tone Arm

Making use of the equivalent diagrams in Figs. 4 and 5, the over-all response characteristic of the system can be calculated. From these figures it can be seen that current i_3 represents the velocity of the ribbon with respect to the insulated plate and current i_1 represents the velocity of the stylus. The ratio of i_3 to i_1 will provide the response characteristic of the pickup with respect to frequency. For high frequencies this can be calculated from

$$\frac{i_3}{i_1} = \left(\frac{C_3}{C_2 + C_3}\right) \times \left(\frac{1}{1 - (2\pi f)^2 L_2 \dfrac{C_2 C_3}{C_2 + C_3}}\right).$$

For low frequencies, the response characteristic may be obtained from

$$\frac{i_3}{i_1} = \left(\frac{C_3}{C_2 + C_3}\right) \times \left(\frac{1}{1 - (2\pi f)^2 L_3 (C_2 + C_3)}\right).$$

It will be noted that two peaks in response occur, one at tone-arm resonance and one at the high-frequency resonance of the pickup moving system. Fig. 8 shows the response characteristic as calculated from the above equations.

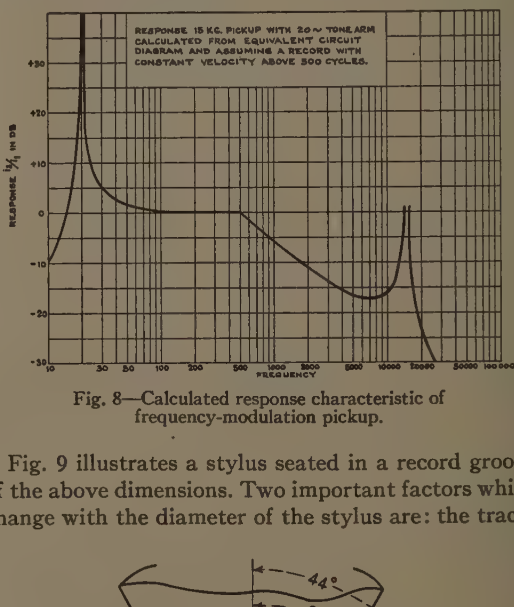

Fig. 8—Calculated response characteristic of frequency-modulation pickup.

(d) Tracking Weight Required to Overcome Vertical Force Due to Lateral Velocity

For proper tracking the stylus must have sufficient vertical force exerted upon it to overcome the vertical component of force due to the lateral velocity of the modulated record groove. Calculations have been made which show the vertical forces exerted upon styli of various radii when seated in a standard groove having an 88-degree included angle, a 0.0023-inch radius cutting stylus and a groove width at the top of 0.0069 inch. In addition to the vertical forces, consideration has also been given to the variations to be expected in pinch effect with different sizes of reproducing styli.

Fig. 9 illustrates a stylus seated in a record groove of the above dimensions. Two important factors which change with the diameter of the stylus are: the track-

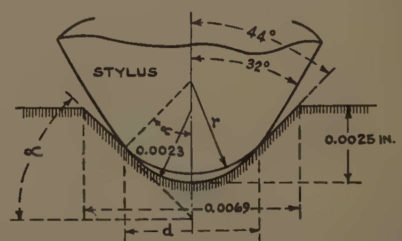

Fig. 9—Tracking diameter and wedging angle of the stylus in the record groove.

ing diameter d and the wedging angle (α). Tracking-diameter d has a direct bearing upon both pinch effect and the high-frequency response and should be kept as small as possible. On the other hand, wedging angle α, which determines the tendency of the stylus to climb the groove wall, should be made as large as possible for

a specified groove. From this it is obvious that a compromise must be made. Fig. 10 shows the variations in d and α with stylus radius, and from observation it can be seen that the stylus radius should not be less than 0.0025 inch or greater than 0.0042 inch. Furthermore, since the curve for angle α is flat from a stylus radius of 0.0025 inch to 0.0042 inch and the curve for diameter d is rising rapidly over this range it appears desirable, when record-groove variations are considered, to use a stylus radius of about 0.003 inch. The importance of the tracking diameter d and the wedging angle α is further emphasized by the curves in Fig. 11. This figure shows, in curve form, the two factors cot α and d tan α which have a direct bearing upon the vertical force due to lateral velocity and pinch effect, respectively. It will be observed that the curve for cot α approaches infinity for a stylus smaller than 0.0023-inch

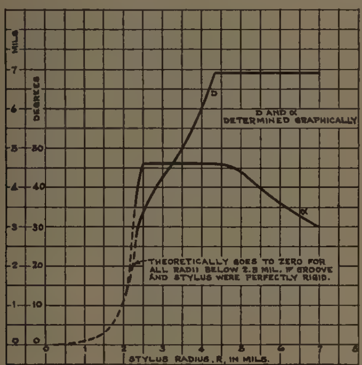

Fig. 10—Variations of tracking diameter and wedging angle for different size styli.

radius, decreases rapidly to about 0.0025-inch radius and then remains essentially flat to 0.0042 inch. At the same time, however, d tan α, the factor governing the pinch effect, increases rapidly from a 0.0023-inch radius to a 0.0042-inch radius and then decreases. Observation of these two curves provides a further confirmation that a stylus radius of 0.003 inch is the best compromise from the standpoint of over-all performance.

In the above equations and those which follow, such factors as the elasticity of record materials and the effective damping of the mechanical system are not included. For this reason the calculated performance characteristics do not provide all the information that might be desired but they do give a general indication of the performance which can be expected. Expressed mathematically the conditions for tracking a laterally cut record exist when

$$E_V = Z_L i_L \cot \alpha$$

where

E_V = minimum vertical force acting on the stylus which will insure proper tracking

Z_L = impedance of stylus and pickup moving system in a lateral direction

i_L = velocity of recording

α = angle at which stylus would ride up groove wall.

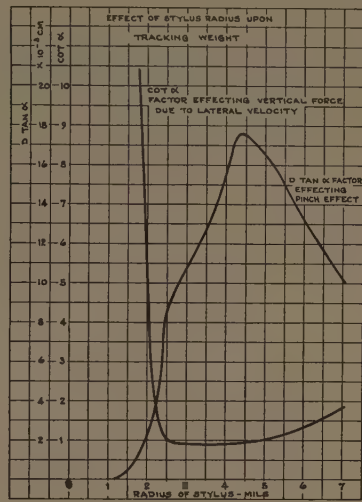

Fig. 11—Effect of stylus radius upon tracking weight.

Utilizing the above equation and the previously determined values for the lateral impedance and the curve for angle α, it is possible to calculate the tracking weight required to overcome the vertical force due to lateral velocity as the stylus radius is varied. Fig. 12 illustrates in curve form the calculated tracking weights for stylus radii of 0.0023, 0.003, 0.004, and 0.005 inch. It will be noted that a stylus radius of 0.003 inch to 0.004 inch requires the least tracking weight.

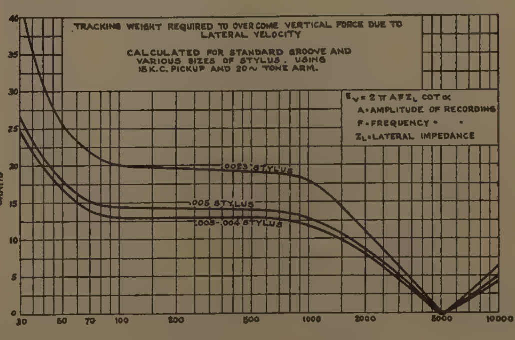

Fig. 12—Calculated tracking weight required to overcome the vertical force due to lateral velocity.

(e) Tracking Weights Required and Relative High-Frequency Outputs to be Obtained with Different Radius Styli

As the stylus radius is increased, it is apparent that for a given groove velocity the output to be obtained at high frequencies will decrease. Calculations have

been made of the expected loss in high frequencies and Fig. 14 shows the curve for this tracking loss at 7000 cycles for styli from 0.0015-inch radius to 0.005-inch radius. It will again be noted that a stylus radius of 0.003 inch is indicated.

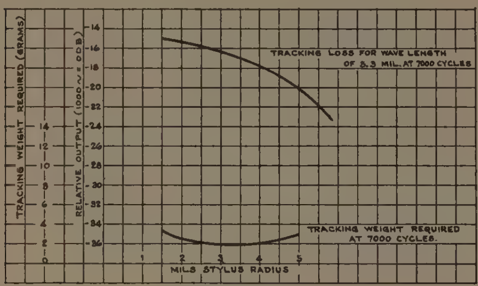

Fig. 13—Calculated tracking weights required and relative high-frequency outputs to be obtained with different radius styli.

2. *Measured Characteristics*

(a) Response Characteristic of Pickup and Tone Arm

Fig. 14 curve (A) shows the over-all response characteristic of the pickup, tone arm, and discriminator as obtained from a frequency record having a 500-cycle crossover point between constant amplitude and constant velocity. The rounded portion of this curve at the crossover frequency is due to the limitations imposed by the electrical network used to provide the recording characteristic. For the purpose of comparison the calculated response characteristic previously shown in Fig. 8 is included in this figure as curve (B).

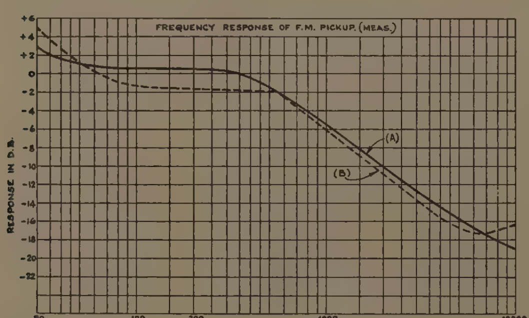

Fig. 14—Measured and calculated response characteristics of frequency-modulation pickup and tone arm.

(b) Tracking Weight Required to Overcome Vertical Force Due to Lateral Velocity

A curve of the measured minimum weight required for satisfactory tracking with an experimental pickup is shown in Fig. 15 as curve (A). Curve (B) gives the calculated values for the same radius stylus. The discrepancy between the curves at the higher frequencies is due to the fact that such factors as pinch effect, elasticity of the record material, and the Q of the mechanical system were not included in the calculations.

(c) Tracking Weights Required and Relative High-Frequency Outputs to be Obtained with Different Radius Styli

Fig. 16 shows the correlation between calculated and measured loss in response at 7000 cycles as the stylus radius is increased. Also shown in the same figure are calculated and measured curves of the tracking weight required for different styli at 7000 cycles. It will be noted that while the actual weights required are somewhat higher than the calculated values, because the

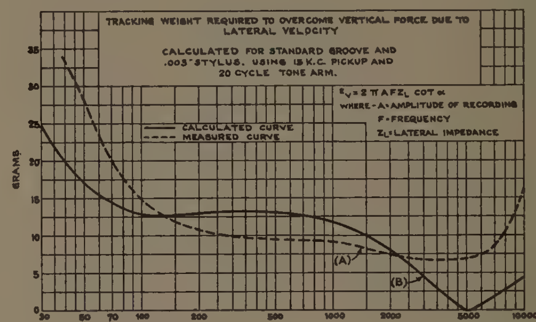

Fig. 15—Measured and calculated tracking weight required by frequency-modulation pickup to overcome vertical force due to lateral velocity.

factors previously mentioned were not included in the calculations, the curve shapes are similar indicating that the choice of a 0.003-inch stylus is a desirable compromise.

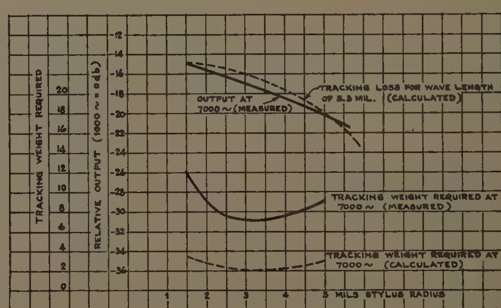

Fig. 16—Measured and calculated tracking weights required and relative high-frequency outputs to be obtained with different radius styli.

(d) Change in Diode Current with Stylus Displacement

Fig. 17 shows the over-all linearity existing between current in the diode resistor and displacement of the stylus. This curve shows the combined effect on linearity of the following factors:

1. Change in the capacitance of the pickup with displacement of the stylus.

2. Change in frequency of the oscillator with change in capacitance of the pickup.
3. Change in output of the frequency-discriminator—rectifier combination with change in frequency.

The departure from linearity of this curve represents a distortion of approximately 2 per cent second harmonic and 0.1 per cent third harmonic. The change in frequency which corresponds to the ± 0.0015-inch displacement of the stylus is $\pm 15{,}000$ cycles.

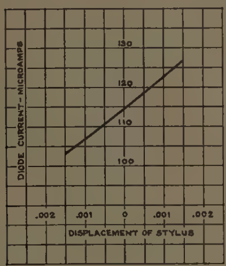

Fig. 17—Change in diode current with stylus displacement.

(e) Comparison of Tracking Weights Required for Various Pickups

Fig. 18 shows the tracking weight required for a frequency-modulation pickup as compared with three other types of phonograph pickups. Curve (A) is a conventional crystal pickup having a normal tracking weight of 70 grams. Curve (B) is a transcription-type magnetic pickup with a normal tracking weight of 45 grams. Curve (C) shows the tracking-weight characteristic of a recently developed crystal pickup which operates at a normal tracking weight of 28 grams and curve (D) shows the tracking-weight characteristic of the frequency-modulation pickup normally operated with a tracking weight of 18 grams.

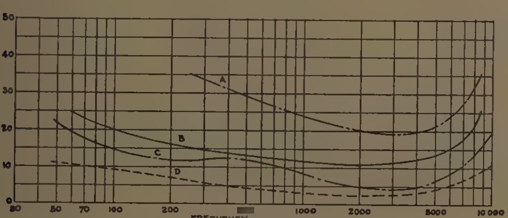

Fig. 18—Comparison of tracking weights required for various pickups.

Conclusions

An experimental frequency-modulation record reproducing system of the type described has been in use for some time. All of the evidence to date indicates that the system is a practicable one and is not adversely affected by changes in temperature, humidity, or line voltage.

The experimental frequency-modulation pickup meets the requirements of a satisfactory pickup to a degree which has not previously been attained in a relatively inexpensive device. The general performance characteristics of a pickup of this type can be calculated within reasonable limits.

From the listeners standpoint, the experimental frequency-modulation phonograph system, which has been described, makes it possible when using conventional shellac records to extend the frequency range of a record reproducing system to 10,000 or 12,000 cycles with an astonishing freedom from surface noise, mechanical noise, and distortion. A further reduction in surface noise can be obtained with shellac records if they are recorded with a high-frequency accentuation characteristic which is comparable to that used in transcriptions. Experimental records of this type have been made. The surface noise obtained from these records with the frequency-modulation reproducing system was reduced to the point where it was not objectionable to the most critical listeners.

Although the calculations and measurements which have been given are confined primarily to 78 revolution-per-minute records the same performance advantages are retained in a frequency-modulation reproducing system designed for transcriptions.

Acknowledgment

The writers wish to acknowledge the valuable assistance of Messrs. H. Belar and R. Snepvangers during the development of the frequency-modulation record reproducing system.

Appendix A

Lateral Mechanical Impedance

The equivalent diagram of the frequency-modulation pickup with a rigid tone arm can be shown as indicated in Fig. 3, where

L_1 = effective mass of needle point
L_2 = effective mass of ribbon and needle referred to point
L_3 = inertia of tone arm and pickup referred to point
C_2 = compliance of needle between point and ribbon
C_3 = compliance of ribbon and needle referred to point
i_1 = velocity of needle point
i_3 = velocity of ribbon with respect to support
Z_L = mechanical impedance in lateral direction

In practice the mass of the tone arm is several hundred thousand times larger than that of either the point or needle. In view of this, further simplification is possible and at high frequencies the equivalent diagram can be set up as shown in Fig. 4.

For these frequencies the lateral mechanical impedance becomes

$$Z_{L_1} = \rho L_1 + \frac{\frac{1}{\rho C_2}\left(\rho L_2 + \frac{1}{\rho C_3}\right)}{\frac{1}{\rho C_2} + \rho L_2 + \frac{1}{\rho C_3}} \quad \text{where} \quad \rho = i\omega$$

$$= \frac{\frac{L_1}{C_2} + \rho^2 L_1 L_2 + \frac{L_1}{C_3} + \frac{L_2}{C_2} + \frac{1}{\rho^2 C_2 C_3}}{\frac{1}{\rho C_2} + \rho L_2 + \frac{1}{\rho C_3}}$$

$$= j\,\frac{-\omega^3 L_1 L_2 + \omega\left(\frac{L_1}{C_2} + \frac{L_1}{C_3} + \frac{L_2}{C_2}\right) - \frac{1}{\omega C_2 C_3}}{\left(\frac{1}{C_2} + \frac{1}{C_3}\right) - \omega^2 L_2}$$

$$= \frac{j}{\omega}\,\frac{\omega^4 L_1 L_2 C_2 C_3 - \omega^2(L_1 C_3 + L_1 C_2 + L_2 C_3) + 1}{\omega^2 L_2 C_2 C_3 - (C_2 + C_3)}.$$

From the above formula it can be seen that infinite impedance occurs when $f = \infty$; $f = 0$; or when

$$\omega = \sqrt{\frac{1}{L_2\left(\frac{C_2 C_3}{C_2 + C_3}\right)}} \quad \text{or} \quad f = \frac{1}{2\pi L_2 \sqrt{\frac{C_2 C_3}{C_2 + C_3}}}.$$

At low frequencies L_1 and L_2 become so small in value that they may be neglected and the equivalent diagram may be as shown in Fig. 5.

The mechanical impedance of the pickup and tone-arm system at low frequencies may be written as

$$Z_{L_3} = \frac{\frac{1}{\rho(C_2 + C_3)} \times \rho L_3}{\frac{1}{\rho(C_2 + C_3)} + \rho L_3} = \frac{\rho L_3}{1 + \rho^2 L_3 (C_2 + C_3)}$$

$$= \frac{j\omega L_3}{1 - \omega^2 L_3 (C_2 + C_3)}.$$

Infinite impedance will then result when

$$\omega = \sqrt{\frac{1}{L_3(C_2 + C_3)}}$$

or when

$$f = \frac{1}{2\pi\sqrt{L_3(C_2 + C_3)}}.$$

This point of infinite impedance occurs when the mass of the tone arm resonates with the total compliance of the pickup, and has been generally given the name of "swinging resonance."

Appendix B

Response Characteristics

Assuming sine-wave motion of the pickup point (perfect tracking) and referring to the equivalent diagram, Fig. 4, it can be seen that the frequency-response characteristic may be calculated from the ratio i_3/i_1, for high frequencies.

Since $i_1 = i_2 + i_3$ and also

$$\frac{i_2}{\rho C_2} = i_3\left(\rho L_2 + \frac{1}{\rho C_3}\right) \quad \text{then} \quad i_2 = i_3\left(\rho^2 L_2 C_2 + \frac{C_2}{C_3}\right).$$

Substituting we obtain

$$i_1 = i_3 + i_3\left(\rho^2 L_2 C_2 + \frac{C_2}{C_3}\right).$$

The ratio of i_3/i_1 becomes

$$\frac{i_3}{i_1} = \frac{1}{1 + \frac{C_2}{C_3} + \rho^2 L_2 C_2}$$

$$= \frac{C_3}{(C_2 + C_3)\left(1 - \omega^2 L_2 \frac{C_2 C_3}{C_2 + C_3}\right)}$$

$$= \left(\frac{C_3}{C_2 + C_3}\right)\left(\frac{1}{1 - (2\pi f)^2 L_2 \frac{C_2 C_3}{C_2 + C_3}}\right).$$

Since the ratio of i_3/i_1, becomes infinite when

$$\omega = \sqrt{\frac{1}{L_2 \frac{C_2 C_3}{C_2 + C_3}}}$$

the frequency of infinite response becomes

$$f_c = \frac{1}{2\pi\sqrt{L_2 \frac{C_2 C_3}{C_2 + C_3}}}.$$

By a similar line of reasoning and making use of the equivalent diagram, Fig. 5, the low-frequency response at resonance may be calculated. In this case

$$\frac{i_3}{i_1} = \frac{C_3}{(C_2 + C_3)(1 - \omega^2 L_3 (C_2 + C_3))}$$

$$= \left(\frac{C_3}{C_2 + C_3}\right)\left(\frac{1}{1 - (2\pi f)^2 L_3 (C_2 + C_3)}\right).$$

Here again the ratio of i_3/i_1 becomes infinite when

$$\omega = \sqrt{\frac{1}{L_3(C_2 + C_3)}}$$

and the frequency of infinite response becomes

$$f_L = \frac{1}{2\pi\sqrt{L_3(C_2 + C_3)}}.$$

Effects of Solar Activity on the Ionosphere and Radio Communications*

H. W. WELLS†, MEMBER, I.R.E.

Summary—The relationships of solar activity on the ionosphere and radio communications may be roughly classified as follows:

(1) There are occasional solar flares or outbursts of ultraviolet light which instantaneously produce radio fade-outs of short duration.

(2) Occasionally solar streams of particles sweep across the earth's orbit producing magnetic storms and auroral displays. The associated ionospheric disturbances may seriously affect radio communication for several days although the effects are more pronounced in polar regions.

(3) The general change of solar ionizing wave radiation in the course of the sunspot cycle governs the average intensity of ionization in the ionosphere. This trend is an important factor governing selection of operating frequencies for radio communication.

General

THE SUN affects the ionosphere and the ionosphere controls radio communications. The sudden radio fade-outs, the occasional complete disruption of radio and wire circuits, the changes of radio-communication conditions from day to night or with season, and the trends from year to year—all of these—may be traced back to the sun, directly or indirectly.

The sudden fade-outs are very spectacular. One moment, communications are normal and the next instant there will be no signals. There is an eruption on the sun, and the fade-out occurs instantaneously. The disappearance of signals applies only to sky-wave communications since line-of-sight transmissions and ground-wave coverage are relatively unaffected. The fade-outs of this type seldom last more than an hour but they have been responsible for many a good radio set being torn apart in futile efforts to locate the trouble.

However, the more severe interruptions to radio communications are associated with magnetic storms and ionospheric disturbances. On such occasions signals may be blanketed out or severely interrupted for several days especially when the wave paths approach polar regions. In addition to the effect on sky-wave communications, strong electrical currents are frequently generated in the surface of the earth which render inoperative many telegraph and other wire circuits. These radio storms are likewise traced back to the sun although the connections, as will be shown later, are not so direct as in the case of the radio fade-out.

Characteristics of the ionosphere are normally recorded by the radio reflection technique developed by Breit and Tuve[1] of the Department of Terrestrial Magnetism, Carnegie Institution of Washington, in 1925. A pulse of short length is sent out from the transmitter and the echo of this pulse is picked up at the receiver. The time difference between the direct (ground) wave and the reflected signal (echo) measures the height of the region of the ionosphere which has returned the signal. Early recordings were on fixed wave frequencies showing the manner with which the height of the so-called "Kennelly-Heaviside layer" changed during the day.

Electromagnetic theory shows that a wave of frequency f, radiated vertically into a medium such as the ionosphere, will penetrate until it encounters sufficient concentration of electrons N to bend the signal around and return it to earth. If N is not adequate, the signal is lost in space. The simplified formula for refraction of the ordinary wave at vertical incidence is

$$N = 1.24f^2 \times 10^{-4} \qquad (1)$$

where N is the electron density in electrons per cubic centimeter and f is the frequency of the reflected wave in megacycles per second.

It will be recalled that the existence of this electrified region surrounding the earth was not publicly accepted until after Marconi's successful communication by radio in 1901 over the curved surface of the earth from England to Newfoundland. It was reasoned independently by Kennelly and Heaviside that radio waves, like light, travel in straight lines, hence signals could travel long distances beyond range of the ground wave only through reflection back to earth from an electrified or conducting region in the outer atmosphere.

Technical literature reveals that as early as 1878—long before the time of Marconi's "wireless"—other scientists had likewise suggested an ionosphere surrounding the earth. Balfour Stewart and Arthur Schuster, however, were studying the earth's magnetism. They observed systematic variations of the earth's magnetism which, they reasoned could be produced by electrical currents circulating in regions of our outer atmosphere.

After Marconi's historic experiment, radio communication over long distances became widely applied. Much practical use was made of this unseen and unidentified "Kennelly-Heaviside" layer. The trends to higher frequencies in the 1920's revealed peculiar developments involving skip distances and phenomenal ranges on low power, even in the daytime.

Concurrently, researches by Breit and Tuve, which were related to studies of the earth's magnetism,

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† Department of Terrestrial Magnetism, Carnegie Institution of Washington, Washington, D. C.

[1] G. Breit and M. A. Tuve, "A radio method of estimating the height of the conducting layer," *Nature*, vol. 116, p. 357; September, 5, 1925.

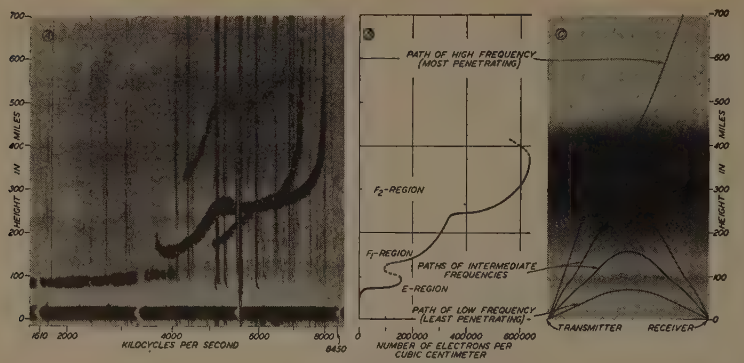


Fig. 1—Exploration of the ionosphere.

(A) Typical record of heights of ion densities obtained with automatic variable-frequency equipment at Kensington, Maryland, 3 P.M., May 15, 1936. (B) Distribution of ions deduced from (A). (C) Density of ionization deduced from (B) showing paths of waves of various frequencies. Diagram (C) shows transmitter and receiver separated for simplication, although in actual work the radio transmitter and receiver are at same station and the wave paths are vertical.

succeeded in directly observing reflections from the outer atmosphere. As mentioned before, the radio "echo" technique developed for this purpose has now

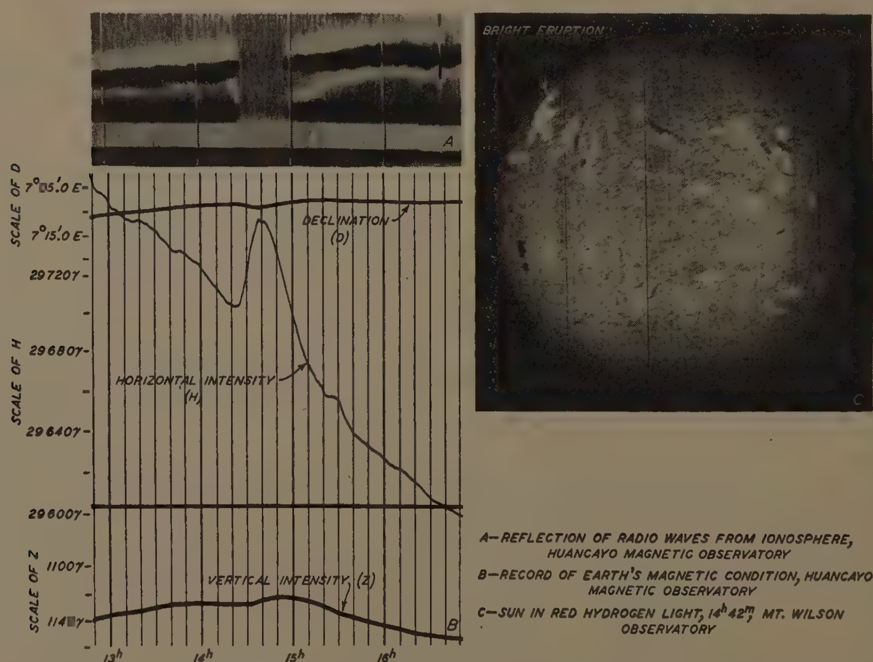


Fig. 2—Changes in earth's magnetism and cessation of radio reflections from ionosphere accompanying bright eruption in solar chromosphere, August 28, 1937 (75 degrees west meridian times).

been widely adopted by investigators in these fields. In England, Appleton and Barnett[2] also succeeded in identifying the ionosphere at about the same time with somewhat different technique. Improved means of exploration of the ionosphere now provide detailed information which is usefully applied to studies of radio wave propagation as well as magnetic variations.

In general, the ionosphere is stratified into at least three separate regions during the daytime. The E region at heights of about 60 miles is the lowest one regularly observed. The F_1 region exists only in the daytime and is found at heights of about 140 miles. The F_2 region is the highest and is found at heights from 200 to 400 miles above the earth. At night the F_1 and F_2 regions merge leaving only the E and the merged F regions.

A brief summary of exploration of the ionosphere is given in Fig. 1. A typical daytime multifrequency recording is illustrated. In the figure, frequencies up to 3500 kilocycles per second were reflected from the E region; frequencies from 3500 to 5200 kilocycles per second were reflected from the F_1 region; and frequencies from 5200 to 7800 kilocycles per second were returned from the F_2 region. Above 7800 kilocycles per second the vertical-incidence signals passed through into space. For oblique-incidence transmissions where transmitter and receiver are separated by real distances, signals on frequencies above 7800 kilocycles per second would be reflected from the ionosphere. For example, the ionospheric conditions

[2] E. V. Appleton and M. A. F. Barnett, "On some direct evidence for downward atmospheric reflection of electric rays," *Proc. Roy. Soc.*, ser. A, vol. 109, pp. 621–641; December, 1925.

illustrated would support communication over a distance of 1000 miles on a wave frequency of 14 megacycles per second.

Long-distance communications may be possible by reflections from any of the ionospheric layers, although the F or F_2 regions are normally most effective in supporting high-frequency signals. Existing methods of translating ionospheric measurements made at vertical incidence into terms of communication frequencies over various distances have been adequately described.[3] Any communications outside the ground range and not in the line of sight depend upon propagation through the ionosphere. Consequently significant changes in ionospheric conditions are reflected as changes in communication conditions. These changes may be for better or for worse. It happens that most of the sudden changes of the ionosphere are associated with poorer communications.

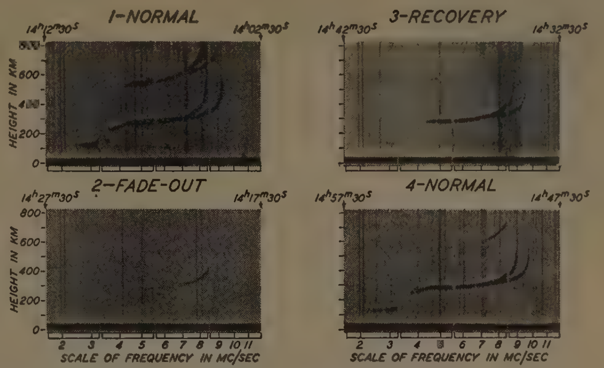


Fig. 3—Radio fade-out recorded at Kensington, Maryland, February 26, 1941, illustrating characteristic sudden development and gradual recovery, 75 degrees west meridian time (E.S.T.) is indicated.

Short-Period Radio Disturbances

The sudden but very temporary ionospheric disturbance generally known as the radio fade-out is very spectacular in its development and effect upon communications, as mentioned above.[4] There is an outburst or eruption on the surface of the sun which is immediately associated with a fade-out of radio signals on the daylight side of the earth only. Communications on the night side of the earth are entirely unaffected. These temporary disturbances are frequently associated with a small pulse or change in the earth's magnetic condition. One very unusual example of this relationship between the sun, the ionosphere, and the earth's magnetism was recorded at the Huancayo Magnetic Observatory on August 28, 1937. The bright eruption illustrated in Fig. 2 occurred near the eastern limb of the sun, and was photographed at about maximum brilliance at 14 hours 42 minutes. The record of radio reflections from the ionosphere on a fixed frequency showed development of a sudden disturbance or fade-out at 14 hours 25 minutes Eastern Standard Time which was followed by complete absorption of all signals until about 15 hours 00 minutes. The recording of horizontal intensity of the earth's magnetic field showed a quiet and normal downward trend until the fade-out at 14 hours 25 minutes at which time the magnetic intensity was suddenly increased. The maximum deviation of magnetic intensity from normal trend was recorded at about the time the solar eruption was photographed which likewise corresponded to the mid-point of the fade-out. After the fade-out the magnetic pulse disappeared and the normal trend of magnetic intensity was continued.

Sudden ionospheric disturbances of this type occur in various degrees of intensity; some weak—some strong.[5] The highest intensities are usually observed when the radiation from the solar eruptions is most direct. Analyses of a number of these effects show that about 60 per cent of the fade-outs endure less than 15 minutes, while an occurrence of more than an hour is very unusual.

Without special facilities, it may be difficult to recognize a radio fade-out until after the disturbance has passed. It is characterized by an abrupt beginning and a somewhat more gradual recovery. The associated solar flare may only be identified by special observing apparatus. It cannot be seen with the naked eye. This effect is contrasted with the disturbance associated with "radio storms" to be discussed later which develop gradually and last much longer. Sudden fade-outs occur at random intervals, sometimes several in the same day and occasionally none at all in a month. In general, their frequency of occurrence follows the sunspot cycle. When sunspot numbers are high, fade-outs are more numerous, and when sunspot numbers are low, fade-outs are less frequent.

The generally accepted explanation of these phenomena which completely absorb radio waves is quite simple. Ultraviolet light of the solar flare penetrates into the lower part of the ionosphere. It is not absorbed extensively in passing down through the normal F_2, F_1, and E regions. However, at some level below the normal E region the intense radiation is absorbed and produces a high degree of ionization. This ionization is in a region of relatively high molecular density, which results in complete absorption and the dissipation of energy of high-frequency radio waves. It is interesting to note that fade-outs occur and disappear without producing any noticeable effect on the other ionospheric regions. In spite of the intense ionization just below the E region, there is no appreciable change

[3] N. Smith, "The relation of radio sky-wave transmission to ionosphere measurements," Proc. I.R.E., vol. 27, pp. 332–347; May, 1939.

[4] J. H. Dellinger, "A new cosmic phenomenon," *Science*, vol. 82, p. 351; October 11, 1935.

[5] L. V. Berkner and H. W. Wells, "Study of radio fade-outs," *Terr. Mag.*, vol. 42, pp. 183–194; June, 1937; and pp. 301–309; September, 1937.

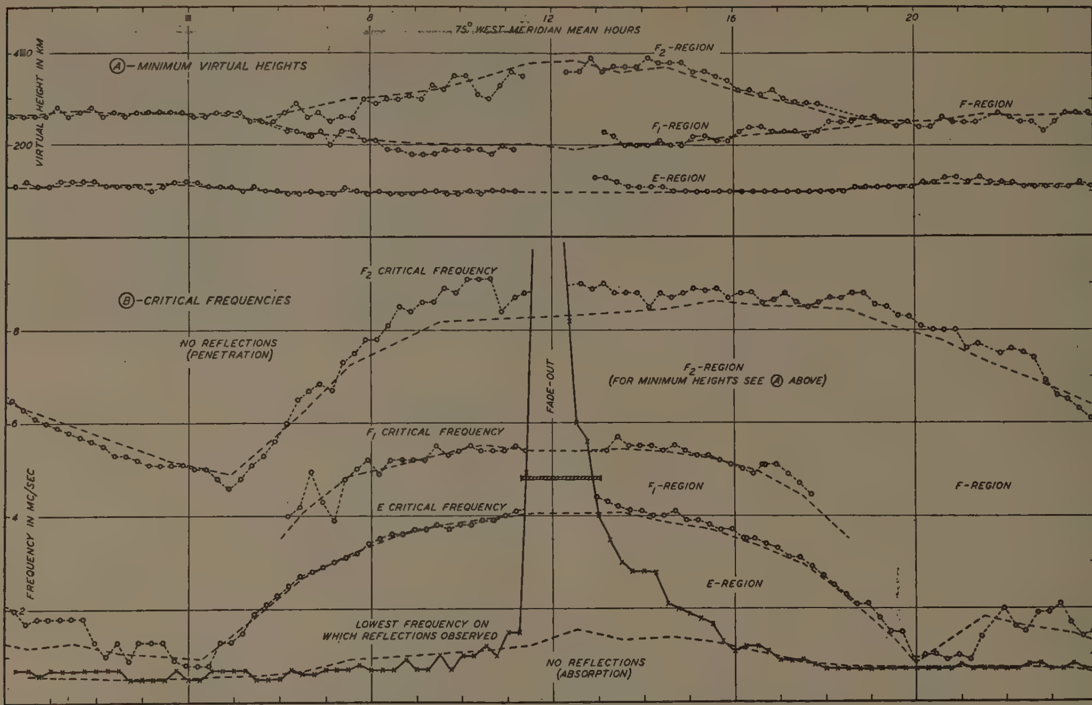


Fig. 4—The ionosphere showing radio fade-out July 31, 1937, determined from automatic multifrequency registrations, Kensington Maryland, (39°01′ North, 77°05′ West).

---o---o---o observed values for o-wave component; – – – – – weekly mean of values for o-wave component (July 26 to 31, 1937); ———— lowest frequency on which reflections observed; |///////| fade-out at Huancayo Magnetic Observatory (12° South, 75° West) determined from automatic fixed-frequency registration (4.8 megacycles per second).

in the degree of ionization in the normal E, F_1, or F_2 layers when it is possible to observe them again following the radio fade-out. Automatic multifrequency recordings of ionospheric characteristics during radio fade-outs have definitely established this fact which is likewise illustrated in Fig. 4. Similarly heights of the several ionospheric regions appear to remain relatively unchanged.

Normal multifrequency recordings of ionospheric characteristics include measurement of the lowest frequency from which reflections at vertical incidence are recorded. In Fig. 4 the lower limit was about 1.0 megacycle per second. Signals below 1.0 megacyle per second in this case were highly attenuated since no reflections were recorded. The radio fade-out, therefore, extends this lowest frequency of reflections up to very high values. When this absorption limit is boosted up high enough it cuts off all reflections from the higher regions. Following the disturbance the recovery to normal is somewhat gradual as illustrated in the figure.

There are certain indications and reports which appear to be substantiated by actual observation that communications on low radio frequencies, say in the range 100 to 200 kilocycles per second or less, may be improved rather than absorbed during the sudden fade-out.[6] The process by which these waves are returned from the ionosphere is more nearly equivalent to

[6] J. E. Best, J. A. Ratcliffe, and M. V. Wilkes, "Experimental investigations of very long waves reflected from the ionosphere," *Proc. Roy. Soc.*, ser. A, vol. 156, pp. 614–633; September, 1936.

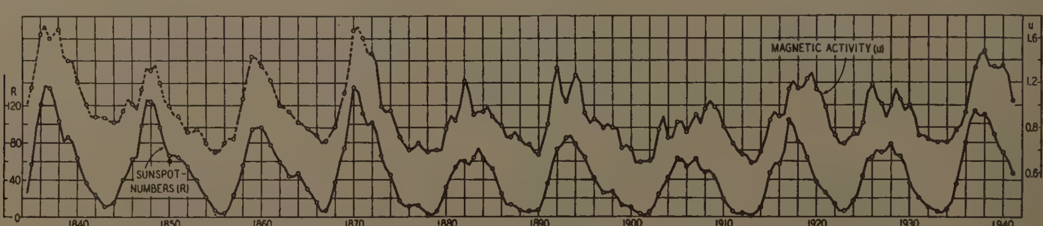


Fig. 5—Geomagnetic activity (u) and relative sunspot numbers (R), annual means, 1835–1941.

that of metallic reflection than to a refractive or bending process. The increased ionization and the improved conductivity in the lower ionosphere resulting from the solar flare could reasonably explain an improvement of signals on the low frequencies.

Long-Period Radio Disturbances

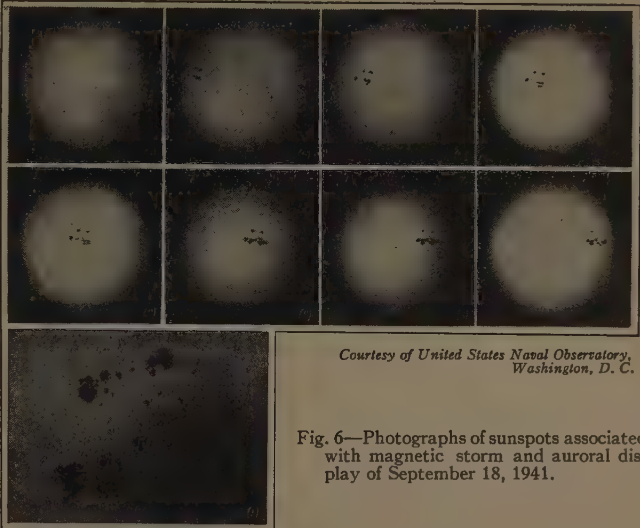
On other occasions, the sun appears to shoot out streams of electrified particles which reach the earth about one to four days later, causing magnegic storms, auroral displays, ionospheric disturbances, and interruptions of radio communications. It will be recalled that the "fade-out" illustrates a *direct* or *instantaneous* relationship between the sun and the earth. Now we have the *delayed* effect which may be very severe and which can disturb communications over the entire world. The intense storms may occur only once or twice a year although moderate disturbances are more frequent. In general these disturbances follow the 11-year sunspot cycle with magnetic and radio disturbances of all types more frequent when the sunspot numbers are high. The comparison of geomagnetic activity with sunspot numbers since 1835 (Fig. 5) emphasizes, however, that there is not a direct, one-to-one relationship between sunspot numbers and geomagnetic activity. Occasions of high magnetic activity and low sunspot numbers, or of low magnetic activity and high sunspot numbers, are frequently recorded. There have even been occurrences of magnetic storms when no sunspots were seen. The general trends, however, are very definitely established.

The intense disturbance of September 18, 1941, is still vividly recalled by many because of the brilliant aurora and the associated effect upon communications. This storm resulted from a very large and active sunspot group. The development of this sunspot group as photographed on eight successive days is illustrated in Fig. 6. It will be noted that on September 12 the group of sunspots was first seen on the eastern limb. The normal rotation of the sun which amounts to one revolution in about 27 days placed the group on the sun's central meridian, pointing at the earth on September 16. The storm, however, did not develop until September 18, two days after the active solar region passed the central meridian. Presumably the solar particles were shot out at the earth on September 16, but did not reach us until two days later.

The North Atlantic circuit disturbance ratings prepared by the Radio Corporation of America, show that communications over the North Atlantic were affected for the best part of three days, namely, September 18, 19, and 20. At the Cheltenham Magnetic Observatory of the United States Coast and Geodetic Survey, the magnetic ranges were the greatest in the history of the Observatory. Auroral displays directly overhead or south of the zenith were seen throughout the United States, and the aurora was reported from points in equatorial regions.

The extent to which radio circuits are interrupted during periods of magnetic activity depends upon the reaction in the ionosphere. Some of the general relationships between magnetic activity and radio communications, especially over the North Atlantic circuits, have been very adequately treated by Hallborg.[7] Such investigations have established the important evidence that circuits over various parts of the world are reacted upon differently by magnetic disturbances. Under the effect of such disturbances the ionosphere exhibits unusual absorption and turbulence. Either of these effects can interrupt normal communications, and a combination of the two is doubly severe.

Absorption during magnetic storms may be equally as intense as that which causes the sudden radio fade-out. Occasions of complete disappearance of signals are recorded. There is increasing evidence to indicate that some, if not all, of the absorption during a magnetic storm occurs in the same general part of the ionosphere where the radio fade-out is produced.



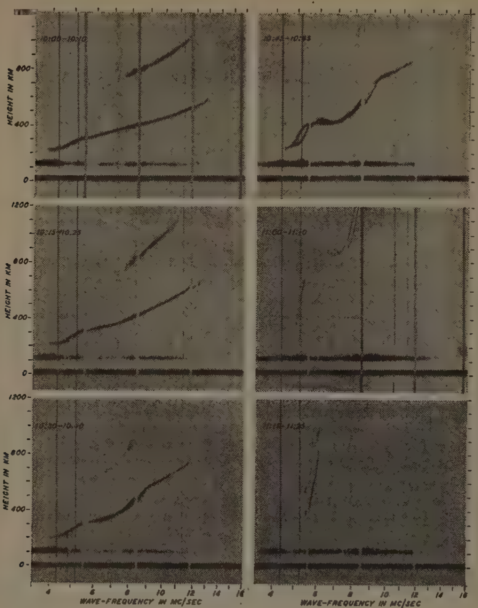
Courtesy of United States Naval Observatory, Washington, D. C.

Fig. 6—Photographs of sunspots associated with magnetic storm and auroral display of September 18, 1941.

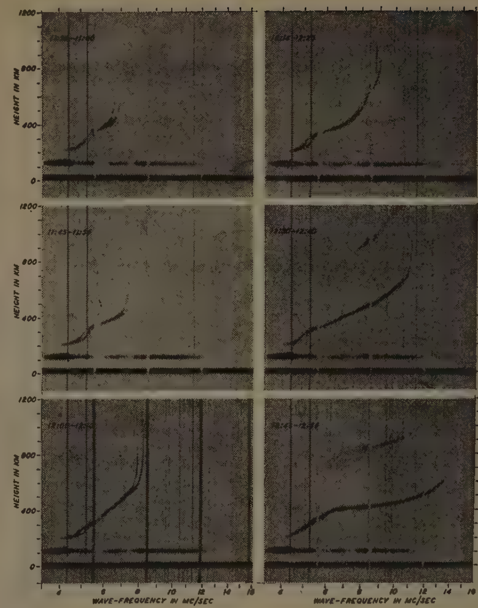
Photograph	Date September	75° West Meridian Time	Photograph	Date September	75° West Meridian Time
		h m s			h m s
(A)	12	11 31 20	(E)	16	10 53 20
(B)	13	10 40 00	(F)	17	13 13 15
(C)	14	11 38 36	(G)	18	15 12 30
(D)	15	12 40 00	(H)	19	11 24 30

(I) Enlarged print of sunspot group, September 17, 1941.

[7] H. E. Hallborg, "Short-wave radio transmission and geomagnetism," RCA Rev., vol. 5, pp. 395–408; April, 1941.



A



B

Fig. 7

A—Ionospheric records illustrating changes during magnetic storm of March 24, 1940, and showing disappearance of F_2 region, Huancayo Magnetic Observatory (12° S, 75° W).

B—Ionospheric records illustrating changes during magnetic storm of March 24, 1940, and showing production of new F_2 region, Huancayo Magnetic Observatory (12° S, 75° W).

Probably the same general explanation holds for both types of absorption; that is, a high degree of ionization is temporarily produced in a region of relatively high molecular density. However, the ionizing mechanisms are distinctive. During magnetic disturbances the absorbing ionization frequently seems to be a downward extension of sporadic E-region ionization. The transition from sporadic E-region ionization to the absorption effect apparently is very critical and involves an effective lowering of height which may not amount to more than a few kilometers.[8]

Turbulence in the ionosphere during magnetic storms is evidenced as abnormally high or low electron concentrations. Systematic analyses of such variation have been made.[9] Probably the most outstanding effect during severe disturbances is the apparent "blowing up" of the outer ionosphere as illustrated in Fig. 7. The F_2 region gradually disintegrates and often completely disappears. Echoes are returned from great heights and the critical frequencies are low indicating greatly reduced electron concentration. Under these circumstances, normal high-frequency sky-wave communication by means of the F_2 layer again is disrupted since the electron concentration is not sufficiently dense to bend the waves back to earth. The lower ionospheric regions, however, are only slightly altered during such disturbances and propagation occasionally takes place in this manner. Wave propagation by the E or F_1 regions, however, is relatively limited as to time of day, frequencies usable, and range of coverage.

As additional evidence of these streams which are shot out from the sun we have actual photographs which sometimes show great extensions or streamers

Courtesy Mount Wilson Observatory, Carnegie Institution of Washington

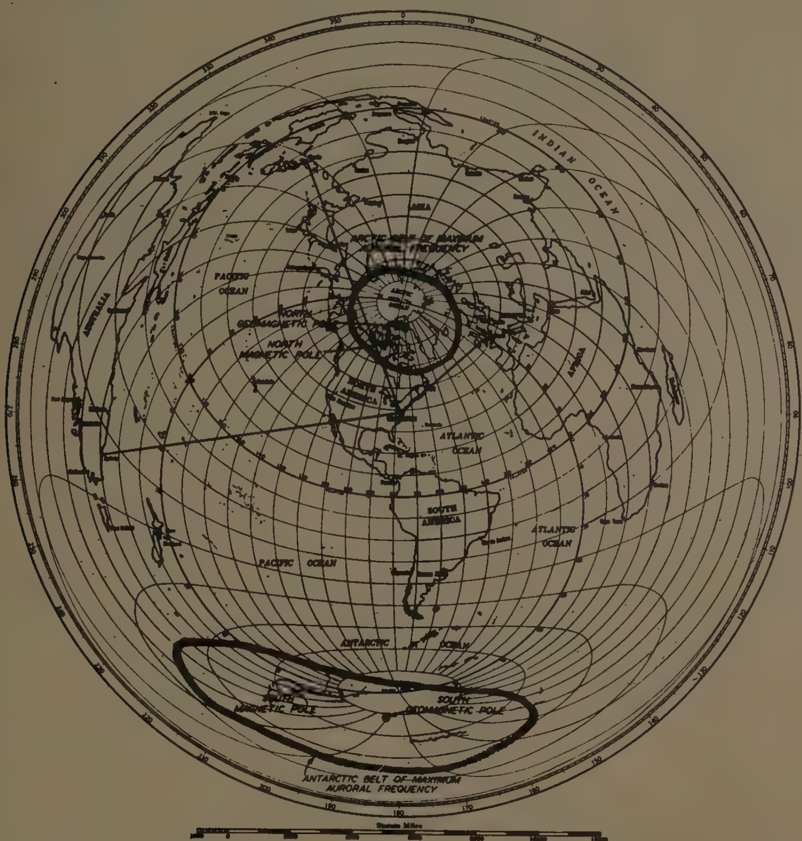
Fig. 8—Streamers over an active sunspot extending into space many thousands of miles.

[8] P. O. Pederson, "The Propagation of Radio Waves Along the Surface of the Earth and in the Atmosphere," Danmarks Naturvidenskábelige, Samfund, Copenhagen, Denmark, 1927.

[9] L. V. Berkner and S. L. Seaton, "Ionospheric changes associated with the magnetic storm of March 24, 1940," *Terr. Mag.*, vol. 45, pp. 393–418; December, 1940.

n particular directions as illustrated in Fig. 8. These streamers occasionally move with high velocity. Some appear to travel right away from the sun while others all back to its surface.

Particles ejected from the sun with a speed of about 1600 kilometers per second would take about 26 hours o travel to the earth's orbit. Evidence has been colected indicating this value to be reasonable for an



Base Map, Courtesy, U. S. Hydrographic Office

Fig. 9—Relationship of great-circle paths from Washington, D. C., with auroral belts.

verage figure although the transit interval varies rom one to three or four days.

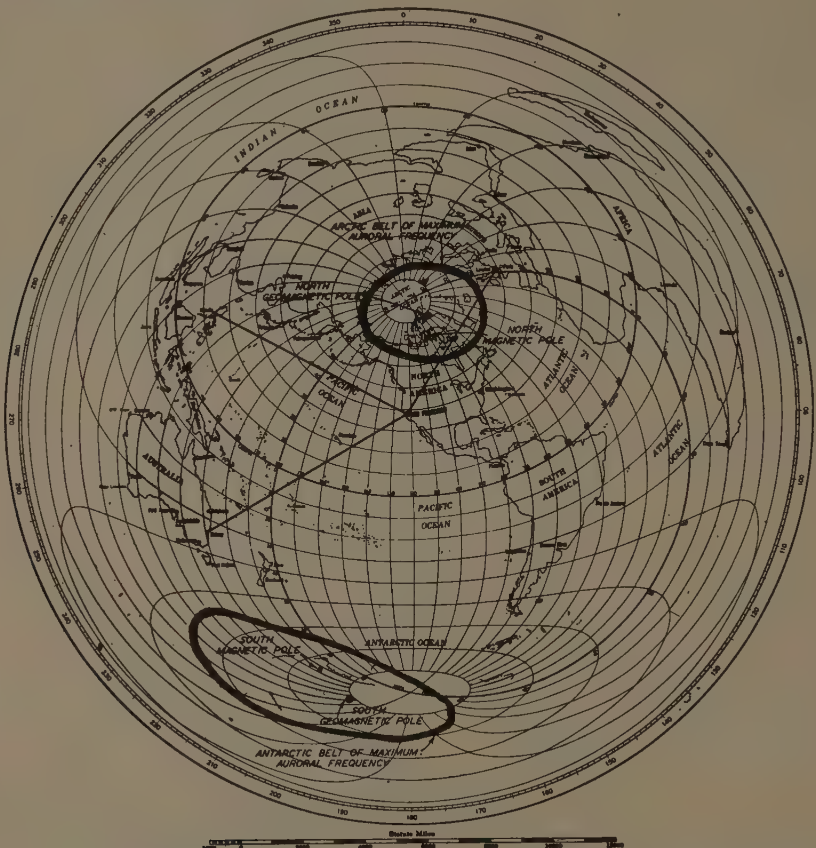
These streams probably are electrically neutral; that , they contain equal quantities of positive and negaive charges. When they approach within the effect of he earth's magnetic field the particles are deflected nto the polar regions producing magnetic disturbances nd auroral displays.[10] It has been shown that the eneral form of magnetic disturbances could be acounted for by a current system in the earth's outer tmosphere. This theoretical current system has high current densities over the polar cap with maximum current flowing in a belt about 23 degrees from the magnetic axis pole. *These theoretical zones of greatest current intensity nearly coincide with the observed zones of maximum auroral activity which have been carefully established by international co-operation.* Outside the auroral zones the theoretical current densities fall off very rapidly.

[10] S. Chapman and J. Bartels, "Geomagnetism," Oxford Uniersity Press, Oxford, England, 1940 ch. 9.

It is common knowledge that radio-communication circuits which traverse the polar regions are erratic in operation and subject to frequent interruptions. Circuits which only approach the polar regions are subject to somewhat less frequent interruptions. The degree of magnetic activity appears to be a factor controlling the disturbances. The North Atlantic circuit from New York to Europe is an excellent example since this radio circuit approaches but does not cross the polar regions. Studies by Hallborg[7] and others of circuit performance over various paths and magnetic activity have indicated unquestionable relationships. They have shown that normal circuit performance is

associated with quiet magnetic conditions, while erratic or poor circuit performance obtains during periods of magnetic disturbance.

It has been mentioned that the normal auroral zones are roughly concentric around the geomagnetic poles. (These are occasionally known as the magnetic axis poles and represent the location the poles would assume if the earth's magnetic field were uniformly distributed.) Furthermore, the auroral activity is an indication of abnormal ionization at levels which include the ionospheric regions. The high current densities required to explain magnetic observations likewise indicate a high conductivity in the outer atmosphere which may be attained through abnormal ionization. One effect of high ionization in the lower ionosphere has already been discussed as the probable cause of the radio fade-out. It seems reasonable to assume that radio-absorbing regions which limit communications over polar areas are very closely identified with the zones of high magnetic and auroral activity. The relationship between auroral zones and communication paths is illustrated by great-circle maps with auroral zones added. Wave paths from the central point to any other point are straight lines. For example, a circuit from Washington to London passes close to the edge of the auroral zone as illustrated in Fig. 9. For magnetically quiet periods this circuit would be expected to operate normally. However, magnetic and auroral disturbances have the effect of expanding the zone radially outward from the geomagnetic poles. Results of magnetic observations in polar regions show that this zone appears to broaden and move toward lower latitudes during periods of intense disturbance so that a station which is ordinarily on the equatorial side of the zone may, during a magnetic storm, be under, or on the polar side of the zone.[10] In this case a very small expansion would sweep the absorbing regions across the North Atlantic circuit path and interfere with operation of the circuit.



Base Map, Courtesy U. S. Hydrographic Office

Fig. 10—Relationship of great-circle paths from San Francisco with auroral belts.

Other circuits from Washington (or the east coast) to the Far East, including China, Philippines, Dutch Indies, etc., cut across the absorbing zones. Commercial as well as amateur experience has established the generally poor nature of communications over these paths. It is also seen that circuits from Washington (or the east coast) to Africa, South America, or Australia

are free from disturbances related to the auroral zones except possibly during severe magnetic storms.

It is interesting to compare radio circuits on great-circle maps based on different points such as Washington and San Francisco in the light of our experience with communications from these points. As mentioned above, the circuit path, Washington to Manila, or to the Far East, traverses a large portion of the auroral zone and is generally unreliable. Now let us examine the circuit, San Francisco to Manila, or to the Far East, as shown in Fig. 10. We see that the wave path falls well outside the normal auroral zone. The dependability of this circuit has been thoroughly established. A relative degree of freedom of interruption from minor magnetic and ionospheric disturbances is assured by the distance between the absorbing zone and the wave path.

For example, a magnetic disturbance which spreads out the absorbing zone just sufficient to interfere with a Washington-to-London circuit probably would not affect operation of a San Francisco-to-Manila circuit (if we had one) because of the greater separation of the latter from the auroral zone. Undoubtedly the complete interpretation of circuit-disturbances is not nearly so simple as suggested above.

We have seen that the zones of maximum auroral and magnetic activity are roughly concentric with the geomagnetic poles. Furthermore, these zones spread out radially from the geomagnetic poles during magnetic disturbances which seem to develop as a result of bombardment of the polar regions by particles from the sun. During slight disturbances the expansion is small, but during severe magnetic storms the zone may extend 20 to 30 degrees beyond normal. The extent of the auroral zone is therefore measured by distances from the geomagnetic poles. Another manner of indicating zones of equal distance from the geomagnetic poles is by a system of geomagnetic co-ordination of latitude and longitude. The zones of maximum auroral activity normally include the geomagnetic latitudes 65 to 70 degrees. Intense magnetic disturbances however will extend these zones out to geomagnetic latitudes 40 to 50 degrees. When this happens radio circuits intercepted by this expanded zone of absorption probably would experience unsatisfactory operation. Radio circuits in the lower latitudes are not so greatly affected during magnetic storms although unusual conditions of absorption and turbulence are experienced. In most cases communication may be maintained although the most satisfactory operating frequencies may be above or below those normally used depending upon the reaction in the ionosphere. Ionospheric recordings near the magnetic equator at Huancayo, Peru, show that the F_2 region of the ionosphere frequently has a lower concentration of electrons during certain phases of a magnetic disturbance. This characteristic is also illustrated in Fig. 7. Occasions, however, of greater than normal electron concentration have also been observed, especially during the preliminary phases of magnetic disturbances.

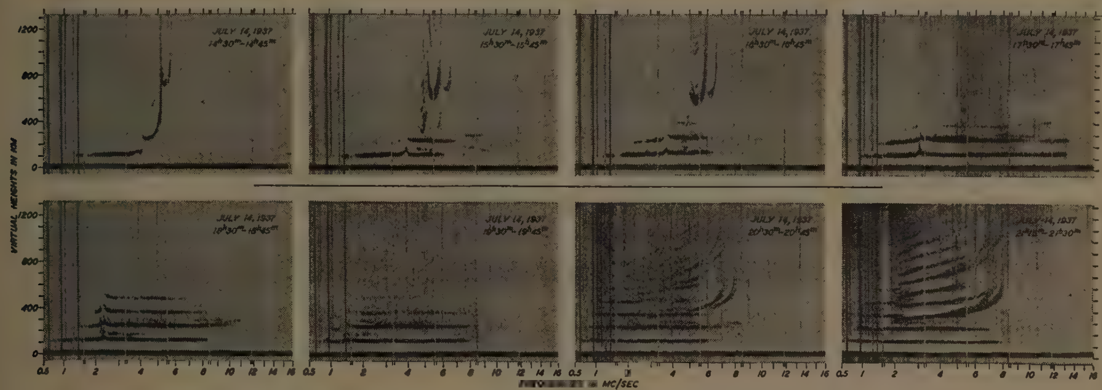


Fig. 11—Ionospheric records, Kensington Experimental Station, during sporadic intense ionization of E region in late afternoon, showing development and subsequent disappearance of the effect.

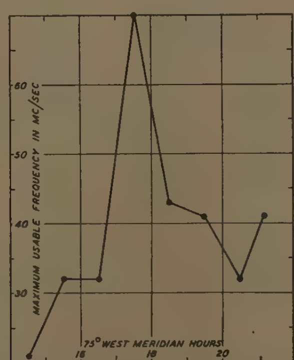


Fig. 12—Maximum usable frequencies over distance of 1000 miles during sporadic E recorded in Fig. 11.

Other Solar Effects on Ionosphere

Solar and magnetic disturbances also seem to be related to the ionospheric effect known as sporadic

E-region ionization. This is a condition which has previously been mentioned as an indication of intense ionization. The stages of development are recorded in Fig. 11. As normally observed, the sporadic E effect is recorded at ionospheric heights slightly higher than the the higher latitudes.[11] These characteristics of distribution resemble auroral occurrences. It is interesting to speculate that subsequent analyses may relate sporadic E as well as aurora more closely to solar corpuscular radiation.

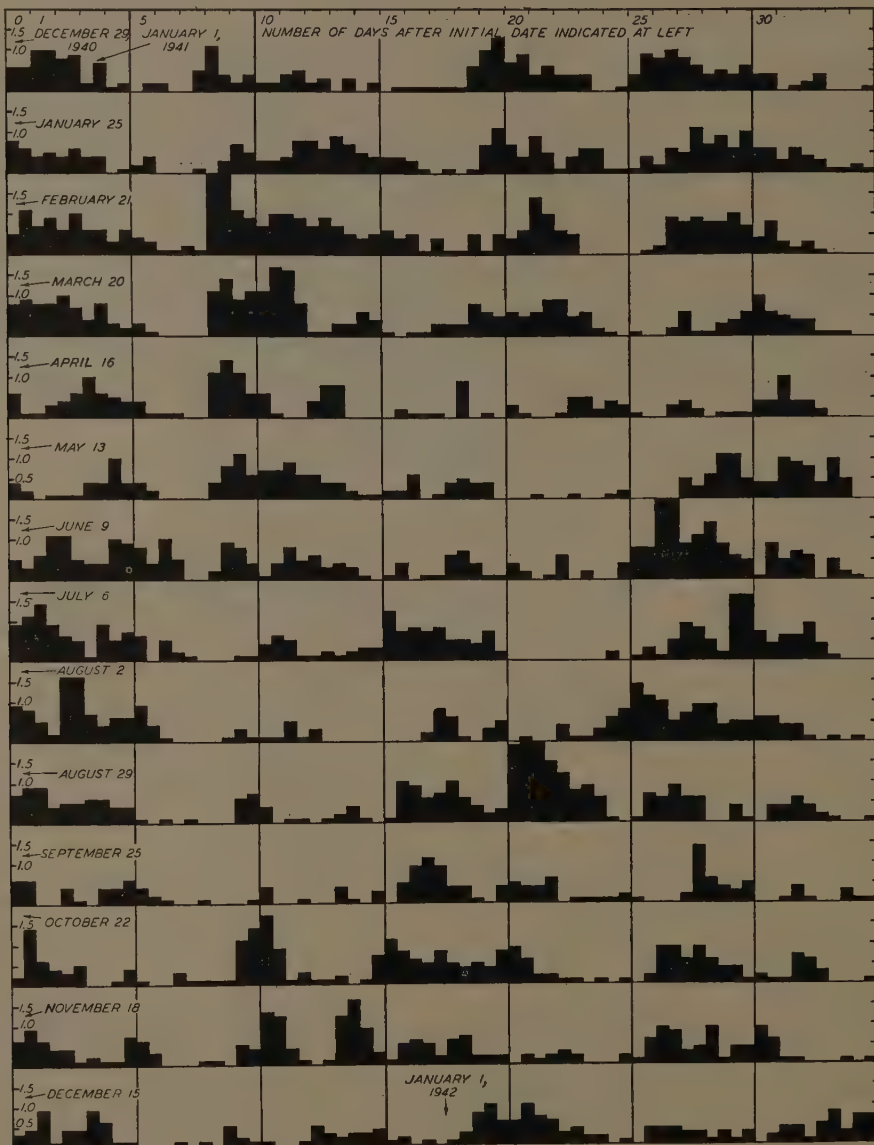


Fig. 13—American character figure C_A, Greenwich half days in 27-day sequences, December 29, 1940, to January 18, 1942.

normal E region. This frequently blankets out or masks off reflections from the higher ionospheric regions. Analyses of the occurrences of this effect from the limited amount of world-wide data available have shown that sporadic E is very seldom recorded near the geomagnetic equator, but that it occurs more frequently at

The sporadic E condition is very effective for communication purposes while it lasts because of the intense ionization and the low height. Occasions have

[11] L. V. Berkner and H. W. Wells, "Abnormal ionization of the E region of the ionosphere," *Terr. Mag.*, vol. 42, pp. 73–76; March, 1937.

frequently been recorded where radio communication on wave frequencies up to 60 and 70 megacycles per second could be supported for communication over distances of 1000 miles or more. The graph of Fig. 12 shows variation of maximum usable frequencies over a distance of 1000 miles during the sporadic E condition recorded in Fig. 11. In this case, a peak of 70 megacyles per second was reached at 17 hours 30 minutes. This maximum did not persist more than about 15 minutes, but wave frequencies exceeding 50 megacycles per second were supported for more than an hour on this occasion. It should be added that communication over distances of 1000 to 1500 miles represents a theoretical limit for single-hop transmission from the E region. Communication over greater distances by such "freak" occurrences is much less probable since reflection from sporadic E effects at two widely separated points in the ionosphere would be involved. There is, of course, the possibility which should not be entirely overlooked that the so-called sporadic E ionization is occasionally more general than at present believed. It is undoubtedly this same effect which makes possible occasional long-distance reception of frequency modulation and other stations on frequencies between 40 and 80 megacycles per second.

One of the factors which again relates the sun with the earth's magnetism and the ionosphere is the occasional tendency for certain types of magnetic and ionospheric disturbances to occur at intervals of 27 days. The sun completes one rotation in 27 days and it is reasonable to assume that active solar areas may be continuously erupting streams of particles which sweep across the earth once in each solar rotation. This 27-day recurrence tendency is illustrated by Fig. 13 which plots magnetic activity in rows of 27 days each. In a figure of this type, January 1, for example, is directly over January 28, and the 27-day interval represents one complete rotation of the sun. Occasions have been noted where moderate magnetic activity has recurred for five or six consecutive solar rotations. However, studies of characteristics of the intense magnetic storms reveal that they seldom recur at all.[12] During periods of high sunspot activity the recurrence tendencies are frequently difficult to determine since the variations are masked by a large amount of random activity. However, during periods of low sunspot numbers these recurrence tendencies are more readily identified because of the absence of other activity which would mask the effect.

The general relationship between sunspot numbers and magnetic storms has already been discussed. Radio observations of ionospheric characteristics for about one sunspot cycle of 11 years are now available. There is a very close correlation between sunspot numbers and average electron density in the ionospheric regions which is revealed by Fig. 14. The curve of electron density at the Huancayo Magnetic Observatory (Peru) for the F_2 region measured at noon shows electron densities in 1938 near sunspot maximum to be more than double the densities recorded in 1933 near sunspot minimum.[12] This marked variation has an important influence upon the selection and application

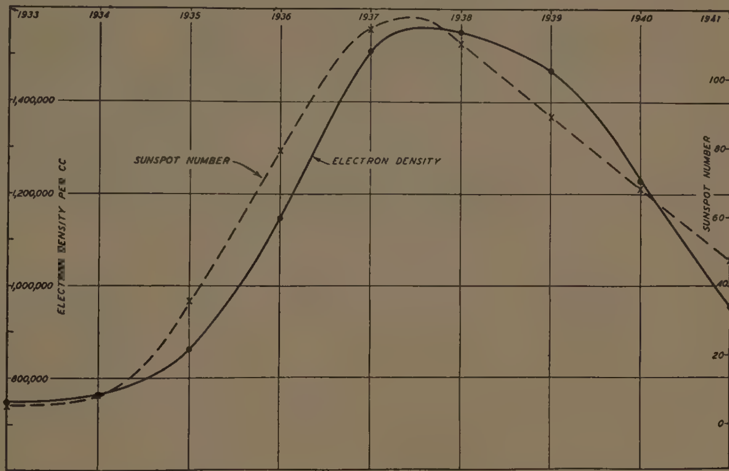


Fig. 14—Comparison of annual average sunspot number with annual average electron density of F_2 region measured at noon at Huancayo Magnetic Observatory, Peru.

of communication frequencies for established communication services. At the present time we are rapidly approaching another period of minimum solar activity and it is to be expected that communication circuits for the next few years will achieve generally satisfactory performance using the same ranges of frequencies which were found suitable about 1931. The time when the low of the sunspot cycle will be reached and the upward trend will start cannot be accurately determined because of frequent irregularities in the length of the sunspot cycle which may range from 9 to 13 years.

In addition to the abnormal effects of solar disturbances on the ionosphere, there are other normal solar effects upon communications. The change from day to night conditions is primarily a solar effect. Similarly the change of ionospheric and communication characteristics with the seasons is primarily a solar effect. Other observations such as those made during solar eclipses add to our knowledge of solar and ionospheric relationships. Such matters, however, are beyond the scope of the present paper.

[12] J. A. Fleming, Eleventh Arthur Lecture of Smithsonian Institution: "The Sun and the Earth's Magnetic Field," presented February 26, 1942. (In press.)

Radio-Frequency-Operated High-Voltage Supplies For Cathode-Ray Tubes*

O. H. SCHADE†, MEMBER, I.R.E.

Summary—**The operation of tuned step-up transformers in self-excited oscillator circuits as high-voltage sources for kinescopes is analyzed. General information and data are given for optimum radio-frequency-transformer design and operating conditions with specified rectifier loads. Practical high-voltage supplies are illustrated ranging from 1 to 50 kilovolts with power-output values of one-quarter watt to 50 watts, respectively. The performance of these supplies in television equipment is discussed.**

Introduction

THE operation of cathode-ray tubes for television requires high-potential direct-current sources, ranging in voltage from less than 1 kilovolt for iconoscopes to 30 kilovolts and higher for projection kinescopes.

The conventional high-voltage supply consists of an iron-core step-up transformer energized from the 60-cycle power line, and a rectifier circuit with smoothing

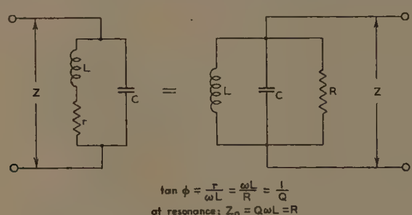


Fig. 1—Power factor and impedance of tuned circuits.

filter. Mechanical and insulation problems make it difficult to construct small 60-cycle transformers with tightly packed windings for voltages exceeding approximately 5 kilovolts. Practical transformers, therefore, are relatively large and heavy and can furnish currents considerably in excess of the usual requirements.

The use of high-frequency-power sources permits a substantial reduction in transformer inductance and results in a relatively simple transformer construction. The input power is generated by vacuum-tube oscillators, which automatically limit the possible power output. This characteristic and the low-energy storage in the small smoothing reactances permit the construction of safe supplies provided the current requirements are not too high.

The theory of tuned step-up transformers points out the necessity of constructing unusually high-impedance secondary circuits to obtain efficient operation. The design of optimum high-voltage coils is, therefore, of prime importance in the construction of practical radio-frequency-operated supplies.

A brief analysis of tuned step-up transformers in self-excited oscillator circuits with rectifier loads will furnish design data for the various circuit components and show their influence on the performance of the high-voltage supply.

The Tuned Step-Up Transformer

The exciting current of a transformer is determined by the reactance of the primary winding and its power factor. The power factor is expressed at radio frequencies by its reciprocal value, the Q value of the reactance. The loss component may be represented as a series resistance r or a shunt resistance R (Fig. 1). For Q values greater than 5,

$$\left.\begin{aligned} r &= \frac{X}{Q} \\ R &= QX \end{aligned}\right\} \begin{aligned} &Q > 5 \text{ and} \\ &X = \omega L \text{ or } X = 1/\omega C. \end{aligned} \tag{1}$$

The magnetizing current of the transformer is canceled with respect to the power source by the operation of tuning the transformer primary. The resonant impedance, hence, of a tuned circuit is

$$Z_0 = R. \tag{2}$$

The secondary of the transformer is tuned by the natural circuit capacitances consisting of distributed coil capacitance, diode capacitance, and stray capacitance. The secondary circuit has, therefore, a natural frequency ω_{02} which determines the operating frequency of the transformer.

A high-voltage radio-frequency transformer is a special case of two coupled tuned circuits. The method of coupling is in general immaterial; the circuit however, must be suitable for stable self-excited oscillations, maintain a substantially constant secondary voltage under considerable external load variations, and load the oscillator efficiently.

The use of critical coupling

$$K_c = 1/\sqrt{Q_I Q_{II}} \tag{3}$$

furnishes a maximum voltage step-up for the no-load condition

$$E_2/E_1 = \sqrt{Z_{II}/Z_I} \tag{4}$$

but it is not suitable for variable loads, because of its dependance on the Q value and impedance of the secondary circuit (equations (3) and (4)). The maximum energy transfer into the secondary circuit is limited to 50 per cent of the power input to the primary circuit.

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† Research and Engineering Department RCA Manufacturing Company, Inc., Harrison, New Jersey.

The stability of the secondary voltage can be greatly improved by increasing the coupling to a value $K \gg K_c$ for a fully loaded circuit. The theoretical efficiency limit then can increase to 100 per cent.

The overcoupled circuit has two coupling frequencies ω_1 and ω_2 which cause a double-hump resonance curve as shown in Fig. 2. The spread of the peaks depends on the coupling,

$$K = \frac{1 - (\omega_1/\omega_2)^2}{1 + (\omega_1/\omega_2)^2}. \tag{5}$$

The relative amplitude of the peaks depends on the relative frequencies ω_{01} and ω_{02} to which the circuits are originally tuned before coupling. It is, hence, possible to control the secondary voltage E_2 by changing the primary tuning without change of coupling or of secondary tuning.

The best voltage stability for variable loads is obtained by operation at the lower coupling frequency, and maximum energy is obtained for a tuning adjustment $\omega_{01} \simeq \omega_{02}$. The latter adjustment, however, is not critical. It is, therefore, the desirable operating condition of the circuit. The voltage step-up is reduced to approximately one half of the maximum obtainable in order to provide high efficiency and good voltage regulation. The latter is in the order of 7 to 15 per cent from no load to full load when the output is measured at the direct-current terminals of practical kinescope supplies and includes oscillator performance. A coupling of $K \geqq 20\ K_c$ is required at full load.

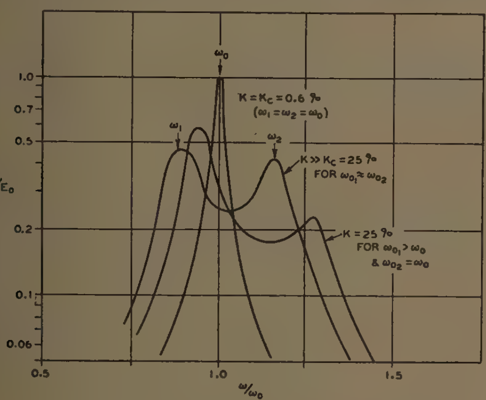


Fig. 2—Frequency characteristics of coupled circuits.

Requirements for Self-Excitation, Inductance, and Q Values

Self-excitation with feedback from the primary winding causes an unstable tuning characteristic as indicated in Fig. 3. A stable oscillation characteristic requires coupling of the grid-circuit inductance L_3 to the secondary circuit L_2 as shown in Fig. 4. The circuit oscillates at the lower frequency peak ω_1 when the winding directions between L_1 and L_3 are as in normal oscillator circuits. Reversal of L_1 or L_3 causes stable oscillation at ω_2.

The full-load Q values of primary and secondary circuits should be high to obtain a large degree of overcoupling ($K \approx 20\ K_c$) with moderate values of K which

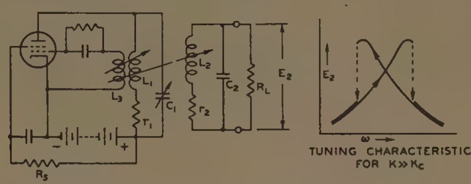


Fig. 3—Oscillator circuit for $K \leqq K_c$, and unsuitable tuning characteristic when $K \gg K_c$.

cannot be made very large because of insulation requirements ($K \approx 25$ per cent).

Desirable values are

$$\left\{\begin{array}{l} Q_I \geqq 10 \text{ when transformer is shunted by the} \\ \qquad \text{reflected plate load } R_p \\ Q_{II} \geqq 20 \text{ when transformer is shunted by the} \\ \qquad \text{equivalent rectifier load } R_L. \end{array}\right. \tag{6}$$

Corresponding inductance values are

$$\begin{array}{l} \omega L_1 \leqq 0.1 R_p \\ \omega L_2 \leqq 0.05 R_L. \end{array} \tag{7}$$

The no-load Q values should of course be considerably higher than the full-load values. A loss of 10 per

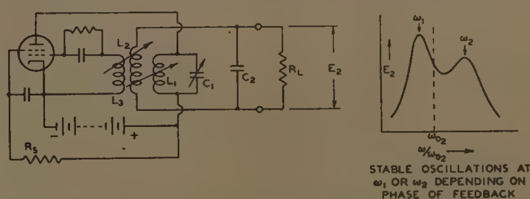


Fig. 4—Oscillator circuit for all values of K, and stable characteristic when $K \gg K_c$.

cent per circuit requires ten times the Q value given in (6); i.e.,

$$Q_{L1} = 100; \qquad Q_{L2} = 200. \tag{8}$$

The Equivalent Rectifier Load

The equivalent rectifier load R_L depends on the rectifier circuit, three types of which are shown in Fig. 5. The alternating-current load R_L is determined by the direct-current load resistance $\overline{R}$, the direct-current output voltage $\overline{E}$, and the alternating peak voltage $\hat{E}_2$ applied to the rectifier tubes.

$$R_L = \frac{(\hat{E}_2)^2 \overline{R}}{2\overline{E}^2} \tag{9}$$

$$R_L = 1/2\bar{R} \text{ for half-wave rectifiers}$$
$$R_L = 1/8\bar{R} \text{ for voltage-doubling circuits.} \tag{9a}$$

The direct-current load resistance $\bar{R}$ of a supply furnishing 1 milliampere at 10 kilovolts is $\bar{R}=10$ megohms. The secondary circuit feeding a half-wave rectifier must, therefore, have an impedance $Z_0=R=10R_L$, i.e., 50 megohms for a secondary loss of 10 per cent. Secondary circuits of such high impedance are too expensive and large for practical use and efficiency is, therefore, sacrificed in favor of size and cost as shown later. Equation (9a) points out the advantage of a doubling circuit, from an efficiency standpoint, because it requires only one fourth the circuit impedance. The circuit of Fig. 5(c) is similar to a doubling circuit, except that D_1 rectifies only part of the coil voltage. The voltage $\hat{E}$ tap is made slightly lower than the desired focusing potential $\bar{E}_1$ for electrostatic types of kinescopes. $\bar{E}_1$ is adjustable by means of the potentiometer P, which allows the addition of B-supply voltage to the radio-frequency voltage. This circuit has a high efficiency, because it does not dissipate power in a bleeder resistance. It maintains also a very stable voltage at increased first-anode current.

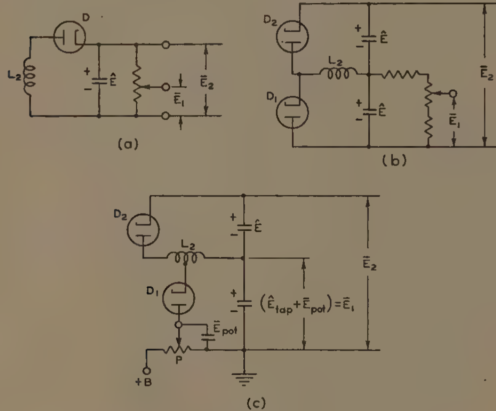


Fig. 5—Rectifier circuits for kinescope supply voltages $\bar{E}_1$ and $\bar{E}_2$.

The Reflected Plate Load

The primary-circuit constants are determined after the secondary coil has been designed from the operating frequency, the total power output $P.O.$ to be supplied by the oscillator, and the oscillator peak voltage swing $\hat{E}_p$. From these, the reflected load,

$$R_p = \frac{(\hat{E}_p)^2}{2P.O.}. \tag{10}$$

The primary reactance ωL_1 is then obtained from (7). The problem of designing an optimum high-voltage coil and determining its operating frequency may be approached in the following manner.

High-Voltage Coil Design

Physical Dimensions

The physical size of the coil depends on the required minimum sparking distances and the power which must be dissipated. The latter is at first unknown. A coil of desirable size for the particular purpose is hence chosen and given a copper cross section consistent with voltage requirements and high-Q values. The winding is subdivided into pies (Fig. 6) with approximately 5 turns per layer, the pie spacing being somewhat less than the pie height in order to maintain the same potential gradient between coils as inside the winding. The coils should be supported by strips of insulating material or by an impregnated paper tubing, which is perforated to permit free circulation of air and to reduce dielectric losses. The coil size indicated in Fig. 6 will dissipate approximately 6.5 watts in a horizontal position.

Design for Optimum Electrical Characteristics

The power loss in the coil is given by

$$P = (E_2)^2/R. \tag{11}$$

The equivalent shunt-loss resistance R at resonance (see Fig. 1 and (1)) can be written $R=L/rC$. Thus,

$$P = (E_2)^2\frac{rC}{L}. \tag{11a}$$

For given values of secondary voltage E_2 and tuning capacitance C, which should be as small as possible, a minimum for the power loss requires a high L/r ratio. At low frequencies, this ratio has a constant value, depending only on the total copper cross section

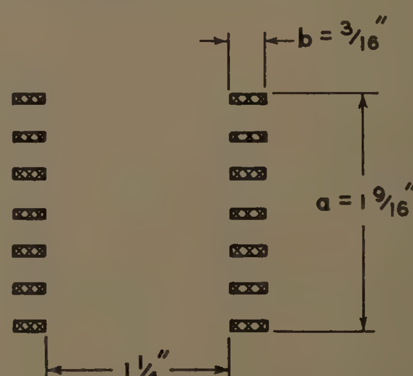


Fig. 6—Dimensions of high-voltage coil for 10 to 15 kilovolts.

of the coil and its shape. At higher frequencies, the coil resistance r increases because of eddy currents, as follows:

$$r = r_0(1 + k^2) \tag{12}$$

where k is the eddy-current factor expressed as

$$k = \frac{0.04N'd^3}{l} f$$

and r_0 = direct-current resistance, ohms
0.04 = constant for particular coil shape
N' = total number of insulated wire strands in cross section of coil
d = strand diameter, inches
l = effective length of coil ($a+b$ in Fig. 6), inches
f = frequency, cycles per second.

It is apparent from (12) that operation at high frequencies requires a small wire or strand diameter d. If it is desired to use Litz wire, we may select No. 41 enamel wire as the smallest desirable wire for strands, but are at liberty to use a single wire or parallel wires (Litz) per turn, thus being able to vary L and f without affecting the copper cross section or any of the remaining factors which determine k in (12). The coil in Fig. 6 contains 4200 strands of No. 41 wire in its cross section; i.e., it may be given as 4200 turns of single No. 41 wire or 2100 turns with 2 parallel strands of No. 41 wire, etc. The tuning capacitance C is estimated to be $C=7$ micromicrofarads (coil capacitance $=3$ micromicrofarads).

The lowest operating frequency of the circuit with $N=4200$ turns ($L=387$ millihenries) is 96 kilocycles, at which the eddy-current factor k^2 has still a negligible value ($k^2=0.037$). The equivalent shunt resistance R has, therefore, the optimum value obtainable with this coil size: $R=22.5$ megohms but the value of Q is only 97. Fig. 7 shows the results of paralleling strands to vary L and f as explained above. The shunt resistance R decreases to 50 per cent of its optimum value at the frequency where Q goes through a maximum.

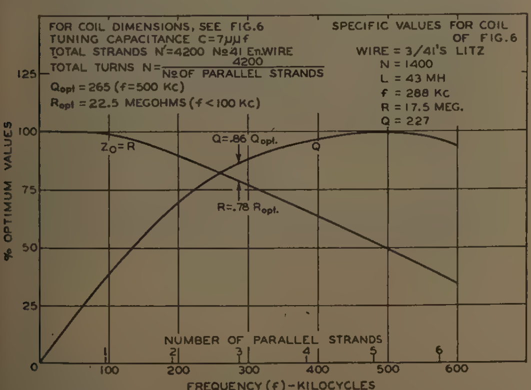
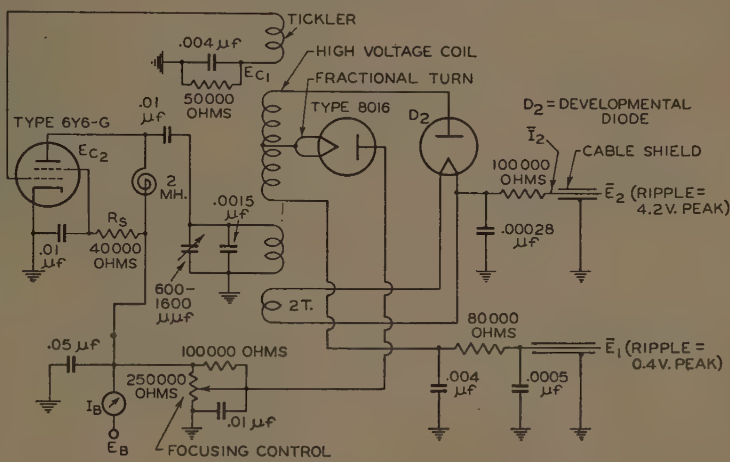


Fig. 7—Efficiency and Q values of coil (Fig. 6) versus number of strands per turn for a fixed product (strands × turns).

A good compromise between efficiency and voltage regulation, which depends on coupling and Q values as explained, indicates 1400 turns of 3-strand Litz wire with $L=43$ millihenries, a resistance $R=17.5$ megohms, $Q=227$, and an operating frequency $f=288$ kilocycles. Other factors, such as winding time, cost of wire, etc., may influence this choice. The maximum peak voltage for $P=6.5$ watts is, hence, $E_{max}=15$ kilovolts for this coil.

Tubes and Circuit Assembly

Efficient operation of the oscillator tubes requires class C excitation and low plate-voltage loss. Beam power tubes such as the 6L6 and 6Y6G are, therefore, especially suitable for use at low supply voltages. The 6Y6G can furnish 15 watts power with 75 to 85 per cent efficiency at voltages between 300 and 375 volts. The grid-leak bias should be $E_{c1}=2E_{cc0}$. The screen-grid voltage is made self-regulating by a series resistance R_s (Fig. 8). It varies from approximately 65 volts at no load to 120 volts at full load and, thus, aids the voltage regulation of the supply. Larger output powers require parallel operation of tubes.



OPERATING CONDITIONS (APPROX.)

E_B VOLTS	I_B MA.	E_{C_2} VOLTS	E_{C_1} VOLTS	$\bar{E}_1$ KV.	$\bar{E}_2$ KV.	I_2 MA.
+325	31	+68	−36	2	10.6	0
+325	62	+115			9.4	0.95

Fig. 8—Circuit and operating conditions of the 10-kilovolt supply for kinescopes.

High-Voltage Rectifier Tubes

Standard high-voltage rectifiers such as the 2X2 or 2V3-G require considerable heater power and are not designed for high-frequency operation. The development of special diodes for rectification of high radio-frequency voltages was therefore indicated. The RCA-8016 requires a cathode power of only one-quarter

watt and thus permits economical radio-frequency heating from the oscillator source.

Smoothing Filter Requirements

The filter capacitances have small values because of the high operating frequency (300 kilocycles for a 10-

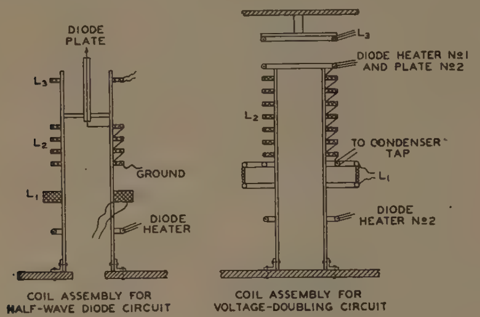


Fig. 9—Typical high-frequency transformer assemblies.

kilovolt supply). In contrast to conditions with 60-cycle operation, the ripple voltage is determined substantially by the ratio of the sum of diode and stray

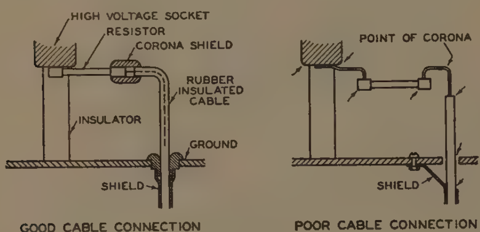


Fig. 10—Corona shielding of cable connections.

capacitances to the filter-condenser capacitance. The actual ripple percentages must be of considerably lower value than in 60-cycle filters to avoid capacitive coupling and interference with receiver operation. Typical values are given in Fig. 8 for a 10-kilovolt kinescope supply.

Circuit Assembly

The particular form of the transformer assembly depends on the type of circuit and the required sparking distances. Typical assemblies are shown in the sketches of Fig. 9. The operation at high radio-frequency voltages emphasizes corona effects because of increased dielectric losses in ionized air. The fine-wire, high-potential ends of transformer windings, must, therefore, be protected against power loss and destructive effects due to corona by guard rings or conductors of sufficient radius of curvature as illustrated in Figs. 9 and 10. This requirement also includes diode terminals and filter circuit.

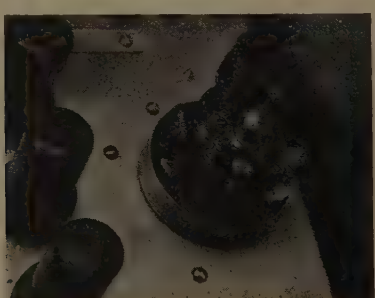


Fig. 11—A 1-kilovolt high-voltage supply for iconoscopes.

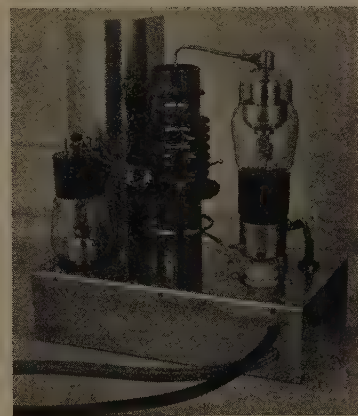


Fig. 12—A 10-kilovolt high-voltage supply for kinescopes.

Developmental Voltage Supplies

A very small voltage supply for iconoscopes is shown in Fig. 11. It was built several years ago for battery operation and is housed in a coil shield $2\frac{1}{2}$ inches in diameter. The 955 oscillator tube takes 8 milliamperes at 180 volts to supply 1 kilovolt to a bleeder circuit

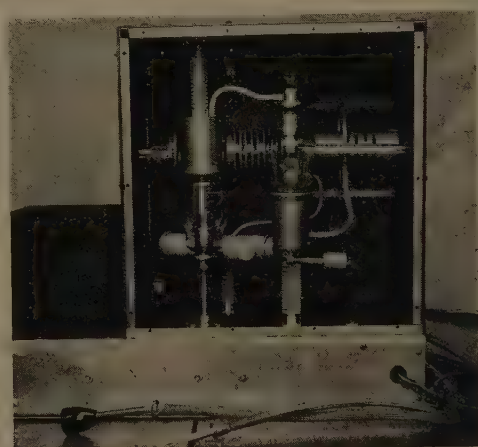


Fig. 13—Arrangement of a 30-kilovolt voltage-doubling circuit.

and iconoscope. The operating frequency is 1.2 megacycles. The small diode is an experimental tube. The larger kinescope supply shown in Fig. 12 operates between 7 kilovolts and 12 kilovolts and measures $7\frac{3}{4} \times 4\frac{1}{4} \times 9$ inches. The supply includes the oscillator tube, which is separated by a heat shield from the transformer assembly. The housing is ventilated at the oscillator but otherwise closed, to prevent dust precipitation on the high-voltage conductors. Operating data are given on the circuit diagram in Fig. 8.

A 30-kilovolt projection-tube supply with separate oscillator for the focusing voltage is shown in Fig. 13. Transformers and rectifier assembly are housed in dust-tight shields. The outside dimensions of the second anode supply are $11 \times 11 \times 12$ inches high. The focusing voltage can be varied from 4 to 7 kilovolts by tuning the primary of its oscillator circuit. The main second-anode supply employs a voltage-doubling circuit energized by three parallel 6Y6G oscillator tubes. Both supplies are operated in series to maintain a desired voltage ratio under varying load conditions. Circuit and performance are shown in Figs. 14 and 15.

A number of radio-frequency-operated supplies for various voltages have given trouble-free service in the laboratory and in television equipment. Voltage stability and focus regulation under

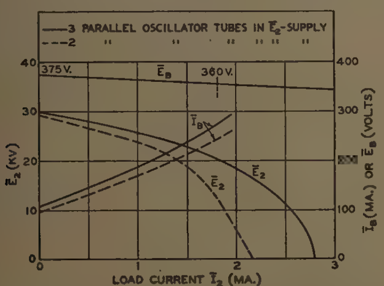


Fig. 15—Regulation characteristics of the 30-kilovolt supply.

actual operating conditions are quite satisfactory. Little difficulty was experienced in preventing oscillator interference with television equipment but isolating resistors or chokes may be required when a single source of filament or B supply is used.

Conclusions

Actual performance has proved that high-voltage supplies energized by vacuum-tube oscillators at high

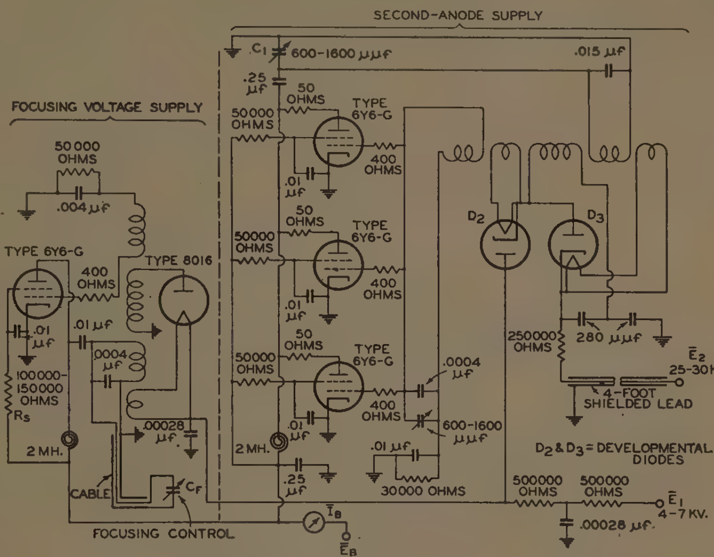


Fig. 14—Circuit of the 30-kilovolt supply for projection kinescopes.

frequencies are practical as kinescope and iconoscope supplies. The obtainable power output is limited by the oscillator power. This method permits the construction of safe supplies, where the current requirements are not too high. This low-power reserve also protects the kinescope and rectifier in case of sparkover or accidental short circuits because of the small short-circuiting current. It must be remembered, however, that currents of dangerous magnitude are obtainable, depending on the voltage step-up and the oscillator power. For such conditions, due precautions for safety must be taken.

The cost of high-frequency-operated supplies compares favorably with 60-cycle supplies when the oscillator power is moderate and when, consequently, small oscillator tubes can be used. Kinescope supplies for voltages up to 30 kilovolts and approximately 50 watts output are in this range.

The Institute desires to present to the readers of the PROCEEDINGS effective material of timely value having major tutorial aspects. In pursuance of this policy, there are included in this issue of the PROCEEDINGS and in subsequent issues, installments covering a portion of the text of a forthcoming volume entitled: "Radio Engineers Handbook." Through the courtesy of its author, Professor Frederick E. Terman (at present Director of the Radio Research Laboratory operating under the supervision of the Office of Scientific Research and Development at Harvard University), and of the publishers, the McGraw-Hill Book Company, New York, N. Y., permission has been granted to the PROCEEDINGS to place this material before our readers. Appreciation of this co-operation is here expressed.

The Editor

Network Theory, Filters, and Equalizers*

FREDERICK E. TERMAN†, FELLOW, I.R.E.

PART I

Summary—**In Part I the fundamental properties of networks are reviewed with particular emphasis on two and four-terminal networks. The use of reactive networks for impedance matching is covered.**

Fundamental network definitions and network theorems are reviewed. The general mesh equations of the network are given, together with their solution for input and transfer impedance. Properties of two-terminal reactive networks are presented in terms of zeros and poles from the viewpoint of Foster's reactance theorem. Methods of synthesizing any impedance realizable by a two-terminal reactive network are given.

Reciprocal impedances are discussed, and the methods of deriving a reciprocal network given.

The general properties of four-terminal networks are reviewed in terms of image impedance and image transfer constant. The alternate presentation on terms of iterative impedance and iterative transfer constant is also covered briefly.

The subjects of impedance matching and insertion loss are considered and formulas are presented for the mismatching factor and for insertion loss.

Properties of four-terminal networks based upon T, π, L, and lattice sections are summarized. The use of reactive T, π, and L networks for impedance matching is reviewed, and charts are presented for designing a matching network to meet any given requirements.

I. Network Definitions

A NETWORK is made up of resistances, inductances, capacitances, and mutual inductances connected together in some manner. The resistances, inductances, mutual inductances, and capacitances involved are termed *network constants* or *parameters*. When these parameters are constant, independent of the current going through them, the network is said to be *linear*.

A typical network is illustrated in Fig. 1. The junctions a, b, c, etc., at which the current can divide are termed *branch points*, and the sections of the network between branch points are termed *branches*. A series of branches that form a complete loop is termed a *mesh*. Examples of meshes are shown by the arrows in Fig. 1, where five separate meshes are designated.

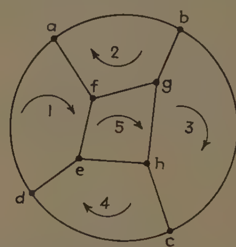


Fig. 1—Network illustrating meshes, branches, and branch points.

A *passive network* is a network containing no source of energy, in which no energy is dissipated other than that accounted for by the resistance elements of the network. An *active network* is a network containing one or more sources of energy, or some sink of energy such as a motor.

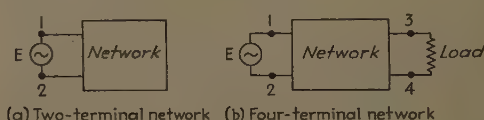


Fig. 2—Two- and four-terminal networks.

The term *two-terminal* is applied to networks operated under the conditions shown in Fig. 2(a). Here the only applied voltage is that shown acting between terminals 1–2 obtained by opening some branch of the network. The term *four-terminal* is applied to a network operating under the conditions illustrated in Fig. 2b. Here a voltage E is applied in series with one branch of the network by opening up this branch to form terminals 1–2, while another branch of the network is opened up to form terminals 3–4, between which an output or load impedance is inserted. A four-

* Decimal classification: R142×R390. Original manuscript received by the Institute, February 17, 1943. Prepublished by permission from "Radio Engineers Handbook", by Frederick E. Terman and the McGraw-Hill Book Company, New York, N. Y.

† Director, Radio Research Laboratory, Harvard University, Cambridge, Massachusetts.

terminal network is equivalent to a transmission system, and is sometimes referred to as a *transducer.*

II. Network Theorems

Superposition Theorem—The current that flows in a linear network, or the potential difference that exists between any two points in such a network, resulting from the simultaneous application of a number of voltages distributed in any manner whatsoever throughout the network is the sum of the component currents at the first point (or the component potential differences between the two points) that would be caused by the individual voltages acting separately.

Reciprocity Theorem—In any network composed of linear impedances, if an electromotive force E applied between two terminals produces a current I at some branch in the network, then the same voltage E acting at the second point in the circuit, will produce the same current I at the first point.

Thévenin's Theorem—Any linear network containing one or more sources of voltage and having two terminals behaves, insofar as a load impedance connected across the terminals is concerned, as though the network and its generators were equivalent to a simple generator having an internal impedance Z and a generated voltage E, where E is the voltage that appears across the terminals when no load impedance is connected and Z is the impedance that is measured between the terminals when all sources of voltage in the network are short-circuited.[1]

Compensation Theorem—If an impedance ΔZ is inserted in a branch of a network, the resulting current increment produced at any point in the network is equal to the current that would be produced at that point by a compensating voltage acting in series with the modified branch, whose value is $-I\Delta Z$, where I is the original current that flowed where the impedance was inserted before the insertion was made.

III. General Mesh Equations of a Network and Their Solution[2]

The response of a network to an applied voltage can be most readily expressed in terms of the mesh currents, i.e., the currents that can be considered as circulating around the closed meshes, as indicated in Fig. 1. The branch currents can then be obtained as the vector sum of the various mesh currents that flow through the branch in question.

Mesh Equations

The voltage and current relations in a network containing n independent meshes can be written as

$$\left.\begin{aligned} Z_{11}I_1 + Z_{12}I_2 + Z_{13}I_3 + \cdots Z_{1n}I_n &= E_1 \\ Z_{21}I_1 + Z_{22}I_2 + Z_{23}I_3 + \cdots Z_{2n}I_n &= E_2 \\ \cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \\ Z_{n1}I_n + Z_{n2}I_n + Z_{n3}I_n + \cdots Z_{nn}I_n &= E_n \end{aligned}\right\}. \qquad (1)$$

In these equations, I_1, I_2, etc., designate the individual mesh currents; E_1, E_2, etc., represent the vector sums of the applied voltages acting around the individual meshes numbered 1, 2, etc. Z_{11}, Z_{22}, Z_{33}, etc., represent the self-impedances of individual meshes, i.e., the impedance around the mesh if all other branches of the network other than those included in the mesh in question were open-circuited. The impedance Z_{12} represents the mutual impedance (coupling) between meshes 1 and 2, as a result of which, a current in mesh 2 produces a voltage drop in mesh 1, and Z_{13} represents the mutual impedance whereby current in mesh 3 produces voltage drop in mesh 1, etc. It is to be noted that reversing the order of subscripts of the Z's does not alter the value of mutual impedance. Thus $Z_{12}=Z_{21}$. Coupling may result either through mutual inductance or from impedance elements common to the two meshes.

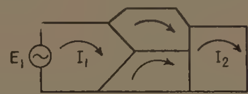


Fig. 3—Network for illustrating meanings of driving point and transfer impedances.

In setting up the system of (1), care must be taken to be consistent in the matter of signs. The positive directions for the mesh currents are assigned arbitrarily. An impedance that is common to two branches is then considered to be a positive mutual impedance when the arrows representing the corresponding mesh current pass through the impedance in the same direction. If the arrows indicate that the corresponding mesh currents pass through the common impedance in opposite directions, then the mutual impedance is the negative of this common impedance. A mutual inductance is positive or negative according to whether it acts with a polarity the same as or opposite to that of a corresponding common inductance.

In setting up a system of relations such as is represented in (1), it is possible to designate the meshes in a variety of ways, subject only to the limitation that each branch of the network must be included in at least one mesh. For example, mesh 1 in Fig. 1, instead of following the configuration *afeda*, could have been defined as *abgfeda*. This would have modified the details in (1), but would have resulted in the same individual branch currents.

In setting up mesh equations and selecting the meshes, there sometimes may be a question as to the number of independent meshes present. In such cases one can use the relation

$$\left.\begin{matrix}\text{number of inde-}\\\text{pendent meshes}\end{matrix}\right\} = \left.\begin{matrix}\text{number of}\\\text{branches}\end{matrix}\right\} - \left.\begin{matrix}\text{number of}\\\text{branch points}\end{matrix}\right\} + 1. \qquad (2)$$

[1] When the sources of energy in the network are constant-current generators, instead of constant-voltage generators, the internal impedance Z is the impedance observed between the terminals when all constant-current generators are open-circuited.

[2] An excellent discussion of the general mesh equations of a network is given by E. A. Guillemin, "Communication Networks," vol. 1, ch. 4, John Wiley and Sons, New York, N. Y., 1931.

A solution of the system of (1) shows that the current I_k in the kth mesh that flows as the result of the voltage E_j acting in the jth mesh is

$$I_k = E_j \frac{B_{jk}}{D}. \tag{3}$$

Here D is the determinant of the system of (1), and is given by

$$D = \begin{vmatrix} Z_{11} & Z_{12} \cdots Z_{1n} \\ Z_{21} & Z_{22} \cdots Z_{2n} \\ \cdot\ \cdot\ \cdot & \cdot\ \cdot\ \cdot\ \cdot\ \cdot \\ Z_{n1} & Z_{n2} \cdots Z_{nn} \end{vmatrix}. \tag{4}$$

Methods of evaluating a determinant (as well as its more important properties) are found in mathematical texts. The quantity B_{jk} in (3) is the principal minor of D and is formed by canceling the jth row and kth column and then moving the remainder together to form a new determinant with one less row and column than D. In evaluating B_{jk}, this new determinant is prefixed with the sign $(-1)^{j+k}$.

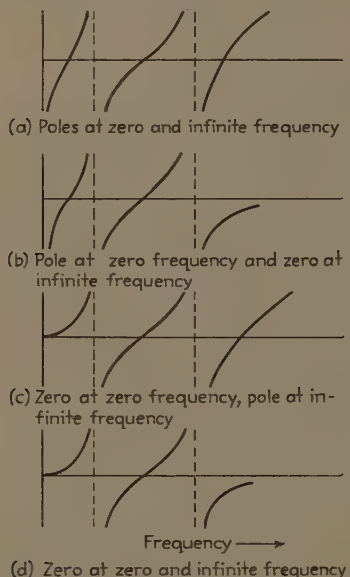


Fig. 4—Reactance as a function of frequency for various classes of reactive networks.

Input and Transfer Impedance

Consider a network having a single applied voltage, with the meshes so arranged that this voltage acts in a branch that is part only of a single mesh, as in Fig. 3. The impedance that the network offers to this applied voltage, i.e., the ratio E_1/I_1, is termed the ***input impedance*** or ***driving-point impedance*** of the network. From (3) this input impedance can be expressed as

$$\text{input impedance} = \frac{D}{B_{11}}. \tag{5}$$

In an analogous manner the ***transfer impedance*** is defined as the ratio of the voltage E_1 applied in mesh 1 to the resulting current I_2 of mesh 2, as indicated in Fig. 3. This transfer impedance can be expressed as

$$\left.\begin{matrix}\text{transfer impedance from} \\ \text{mesh 1 to mesh 2}\end{matrix}\right\} = \frac{D}{B_{12}} \tag{6}$$

where the meshes are so selected that I_2 is observed in a branch that is in only the second mesh of the network and the branch in which the voltage is applied is contained only in the first mesh of the system.

IV. Two-terminal Reactive Networks

The characteristics of a two-terminal network composed of ideal reactances having zero losses are important, because such networks approximate very closely the reactive arms that are used to build up filters, impedance-matching networks, etc.

Foster's Reactance Theorem[3,4]

The driving-point impedance of a two-terminal reactive network behaves as shown in Fig. 4. The impedance curve consists of segments going from minus infinity to plus infinity (except possibly at zero and infinite frequency where a segment may start or stop, respectively, at zero impedance). The slope of the curve is everywhere positive, and is greater than the slope of a straight line drawn to the origin. The frequencies at which the impedance is infinity are termed ***poles***, and the frequencies at which the impedance is zero are termed ***zeros***.

Foster has shown that the driving-point (i. e., input) impedance of a reactive network is uniquely specified by the location of the *internal* zeros and poles, plus one additional piece of information.[5] Expressed analytically, the reactance function can be written as follows:

For a pole at the origin (Figs. 4(a) and 4(b)

$$\left.\begin{matrix}\text{driving-point} \\ \text{impedance}\end{matrix}\right\} = Z$$

$$= \pm j \frac{H}{\omega} \frac{(\omega^2 - \omega_1^2)(\omega^2 - \omega_3^2) \cdots (\omega^2 - \omega_p^2)}{(\omega^2 - \omega_2^2)(\omega^2 - \omega_4^2) \cdots (\omega^2 - \omega_q^2)}. \tag{7a}$$

For a zero at the origin (Figs. 4(c) and 4(d)

$$\left.\begin{matrix}\text{driving-point} \\ \text{impedance}\end{matrix}\right\} = Z$$

$$= \pm j\omega H \frac{(\omega^2 - \omega_1^2)(\omega^2 - \omega_3^2) \cdots (\omega^2 - \omega_p^2)}{(\omega^2 - \omega_2^2)(\omega^2 - \omega_4^2) \cdots (\omega^2 - \omega_q^2)} \tag{7b}$$

where the angular velocities $\omega_1, \omega_3, \cdots, \omega_p$ designated by odd subscripts correspond to the internal zeros of the reactance function, and the angular velocities $\omega_2, \omega_4, \cdots, \omega_q$ designated by even subscripts, correspond to the internal poles of the reactance (see Fig. 4). The plus sign applies when there is a pole at infinite

[3] Ronald M. Foster, "A reactance theorem," *Bell. Sys. Tech. Jour.*, vol. 3, p. 259; April, 1924.

[4] E. A. Guillemin, "Communication Networks," vol. 2, John Wiley and Sons, New York, N. Y., 1935.

[5] Poles or zeros at the origin, or at infinity, are referred to as external, and play no part in the specification of the reactance function.

frequency, while the minus sign applies with a zero at infinite frequency. The sum of the number of poles and number of zeros is one less than the number of independent meshes of the network. The quantity H is a positive real constant that takes into account the fact that one additional piece of information is required to complete the specification of the reactance function.

Foster's reactance theorem shows that the impedance characteristics obtainable from a physically realizable reactance network are quite restricted. This is important, because it limits the characteristic obtainable from filters and other networks.

Synthesis of Reactive Two-terminal Networks

Any driving-point reactance characteristic that can be obtained from any conceivable two-terminal reactive network can be realized by either one of the two networks shown in Fig. 5. The first of these consists of parallel resonant circuits connected in series, with one parallel resonant circuit corresponding to each internal pole. The series condenser C_0 is omitted in the event that the network has a zero at the origin, while the series inductance L_{q+2} is omitted if the network has a zero at infinity. The magnitudes of the circuit components required in the equivalent network of Fig. 5(a) to realize a desired impedance characteristic are given by the following relations:

$$C_k = \left| \frac{j\omega_k}{Z_k} \right| \qquad (k = 2, 4, \cdots, q) \tag{8}$$

where Z_k is the quantity obtained by omitting the term $(\omega^2 - \omega_k^2)$ from the denominator of the corresponding expression (7) for Z and evaluating the modified expression for Z with $\omega = \omega_k$. Corresponding to each C_k, one has

$$L_k = \frac{1}{\omega_k^2 C_k} \cdot \tag{9}$$

If the network has a pole at infinity, then

$$L_{q+2} = H. \tag{10}$$

If the network has a pole at zero frequency

$$C_0 = \left| \frac{1}{Z_0} \right| \tag{11}$$

where Z_0 is the quantity obtained by omitting the ω in the denominator of (7a) under H and evaluating the modified expression for Z with $\omega = 0$.

An alternative method of synthesizing any desired impedance characteristic is to use the arrangement of Fig. 5(b), in which the two-terminal network is built up of a number of series-resonant circuits connected in parallel, with one resonant circuit for each internal zero. In the event that the poles and zeros are so arranged that the network has a pole at the origin, inductance L_0 is omitted. Similarly, if there is a pole at infinite frequency, capacitance C_{p+2} is omitted. The values of the elements in the circuit of Fig. 5(b) required to give a reactance function corresponding to specified zeros, poles, and the given value of H, is obtained from the following equation which can be deduced by a partial fraction expansion of $1/Z_{11}$ as given by (7):

$$L_k = | j\omega_k Z_k' |, \qquad (k = 1, 3, \cdots, p) \tag{12}$$

where Z_k' is the quantity obtained by omitting $(\omega^2 - \omega_k^2)$ from the numerator of expression (7) for

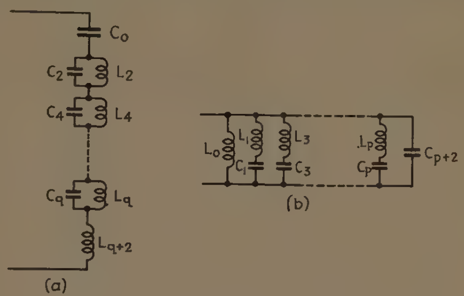


Fig. 5—General forms of reactive networks.

Z and evaluating the modified expression for $\omega = \omega_k$. Corresponding to each L_k, one has

$$C_k = \frac{1}{\omega^2 L_k} \cdot \tag{13}$$

If the network has a zero at infinite frequency

$$C_{p+2} = \frac{1}{H} \cdot \tag{14}$$

If the network has a zero at zero frequency

$$L_0 = Z_0' \tag{15}$$

where Z_0' is the quantity obtained by omitting the ω that multiplies H in (7b) and then evaluating the modified expression for Z with $\omega = 0$.

The networks of Fig. 5 represent networks having the least possible number of reactive elements that can be used to realize a specified impedance characteristic. The same impedance characteristic may also be realized by many other networks other than the two shown,[6] but alternative arrangements will in most cases have additional circuit elements that are superfluous. The least number of circuit elements required is one more than the sum of the internal poles and zeros. One, or even more than one, element beyond this minimum is possible for a given number of internal zeros and poles, but such networks have exactly the same impedance characteristic as the simplest

[6] Ladder networks starting with either a series impedance or a shunt impedance also can be used to develop a specified two-terminal impedance utilizing the minimum possible number of circuit elements. The circuit constants for such networks are obtained from continued fraction expansion of the reactance function. For further information, see pages 198–207 of footnote reference 2.

forms, such as given in Fig. 5, which are called fundamental or canonic forms because they have the least possible number of elements.

V. Inverse or Reciprocal Impedances[7,8]

Two impedances Z_1 and Z_2 are said to be reciprocal with respect to an impedance Z if they are so related as to satisfy the relation

$$Z_1 Z_2 = Z^2. \tag{16}$$

Under practical conditions where reciprocal imped-

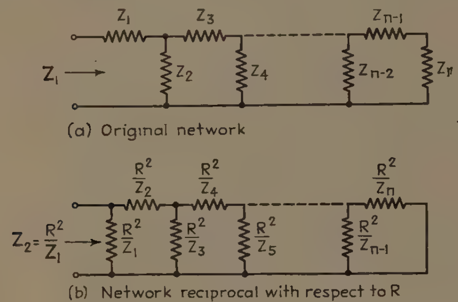


Fig. 6—Reciprocal ladder networks.

ances are of importance, the impedance Z in (16) is always a resistance.

The process of deriving a reciprocal impedance from

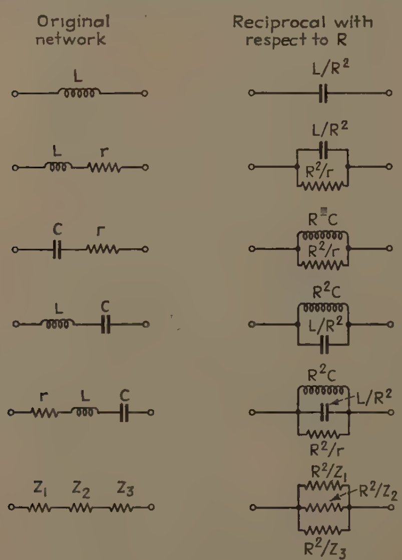


Fig. 7—Basic examples of reciprocation.

a given impedance is termed *reciprocation*. Reciprocation of a ladder network consisting of alternate series and shunt arms, as in Fig. 6(a), leads to a corresponding ladder network of alternate shunt and series arms, as shown in Fig. 6(b).

[7] See p. 204 of footnote reference 2.
[8] A. C. Bartlett, "Theory of Electrical Artificial Lines and Filters," John Wiley and Sons, New York, N. Y., 1931, pp. 53–58.

Various special cases of reciprocation are shown in Fig. 7. These assume that reciprocation is with respect to a resistance R, i.e., Z in (16) is taken as R. It will be noted that the reciprocal of a number of impedance elements in series consists of a number of impedance elements in shunt, with each shunt element being the reciprocal of one of the series elements, and vice versa. It is also to be noted that when two reactive networks are reciprocal, the poles of one coincide with the zeros of the other impedance, and vice versa.

VI. Fundamental Relations Existing in Four-terminal Networks[9]

Methods of Expressing Network Characteristics

Insofar as the four terminals are concerned, the properties of a four-terminal network at any one fre-

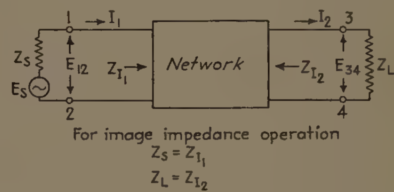


Fig. 8—Four-terminal network operated with terminal impedances.

quency can be expressed in terms of any three independent properties of the network, irrespective of how complicated the network is. Although there are an unlimited number of ways in which three independent constants can be defined, the ones most commonly employed in communication networks are (1) image impedances and image transfer constant; (2) open- and short-circuit impedances; and (3) iterative impedances and iterative transfer constant.

Four-terminal Network Behavior Expressed in Terms of Image Impedances

A network is said to be operated under image-impedance conditions when the internal impedance Z_s of the source of power acting on the input terminals of the network, and the load impedance Z_L at the output terminals, are so related to the network that the impedance looking into the network from the terminals 1–2 with the load connected (see Fig. 8) is equal to the generator impedance, and, similarly, so that the impedance looking into the network at terminals 3–4 with the generator connected equals the load impedance. The generator and load impedances required to produce this condition are properties of the network, and are termed the *image impedances*. They can be designated by the symbol Z_{I_1} and Z_{I_2} for the input and output terminals, respectively (see Fig. 8).

[9] Additional information on these subjects is given in the following books: K. S. Johnson, "Transmission Circuits for Telephonic Communication,"; T. E. Shea, "Transmission Networks and Wave Filters," D. Van Nostrand Company, New York, N. Y., 1929; Guillemin, footnote reference 2.

In dealing with image impedances, the third independent property of the network required to finish the specification of the network behavior is taken as the *image transfer constant* θ, which is defined in terms of the relations

$$\frac{E_{34}}{E_{12}} = \sqrt{\frac{Z_{I_2}}{Z_{I_1}}}\,\epsilon^{-\theta}, \qquad \frac{I_2}{I_1} = \sqrt{\frac{Z_{I_1}}{Z_{I_2}}}\,\epsilon^{-\theta}. \tag{17}$$

The notation is illustrated in Fig. 8. The image transfer constant θ has the same value irrespective of the direction of transmission of energy through the network.

The three network parameters Z_{I_1}, Z_{I_2}, and θ can be defined in terms of the network determinant and its minors, and also in terms of the open- and short-circuit impedance of the network, according to the equations

$$Z_{I1} = \sqrt{\frac{DB_{22}}{B_{11}B_{1122}}} = \sqrt{Z_{oc}Z_{sc}} \tag{18}$$

$$Z_{I2} = \sqrt{\frac{DB_{11}}{B_{22}B_{1122}}} = \sqrt{Z_{oc}'Z_{sc}'} \tag{19}$$

$$\tanh\theta = \sqrt{\frac{Z_{sc}}{Z_{oc}}} = \sqrt{\frac{Z_{sc}'}{Z_{oc}'}} = \sqrt{\frac{DB_{1122}}{B_{11}B_{22}}} \tag{20}$$

where Z_{oc} and Z_{sc} are the impedances at terminals 1–2 with terminals 3–4 open- and short-circuited, respectively, Z_{oc}' and Z_{sc}' are the impedances at 3–4 with 1–2 alternately open- and short-circuited, while D is the determinant for the network formed by short-circuiting terminals 1–2 and 3–4. The B's are minors of this determinant, with B_{1122} being the minor formed by striking out both first and second rows and first and second columns of the determinant.

Image-impedance operation of a network can be conveniently related to the behavior of a transmission line expressed in terms of wave trains. Although wave trains obviously cannot exist in a network having lumped constants, it is nevertheless frequently convenient to explain the behavior of a four-terminal system, insofar as the terminals are concerned, in terms of wave trains just as though these wave trains actually existed within the network instead of being hypothetical. The image impedances correspond to the characteristic impedance of the transmission line, but unless the network is symmetrical about its mid-point, there will be two image impedances because of the fact that the network can be considered as equivalent to a transmission line that is unsymmetrical and so has an impedance transforming action upon a wave train. The image transfer constant θ of the network is likewise analogous to the hyperbolic angle of the transmission line. The real part of the image transfer constant is called the *attenuation constant*, and can be considered as causing the hypothetical wave train in the network to be attenuated in magnitude. The imaginary part of the image transfer constant corresponds to the phase constant of the transmission-line hyperbolic angle, and causes a shift in phase of the hypothetical wave train. If the load impedance does not equal the image impedance on the output side of the network, the effect on the voltage and current relations is as though a wave train existed in the network, and was reflected by the load impedance, just as in the case of a transmission line.

When several networks are connected together in cascade on an image-impedance basis, as illustrated in Fig. 9, then the image impedances at the input and output terminals are the image impedances at the input terminals of the first network and the output terminals of the network, respectively (Z_{I_1} and Z_{I_2}'' in Fig. 9). The image transfer constant of such a system is

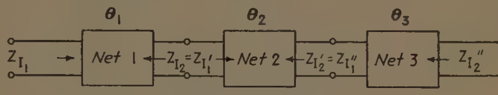


Fig. 9—Networks connected in cascade on an image-impedance basis.

$$\theta + \theta_1 + \theta_2 + \cdots + \theta_m \tag{21}$$

where θ_1, θ_2, etc., represent the image transfer constants of the first, second, etc., component networks of the system.

In a system consisting of a number of networks connected in cascade on an image-impedance basis, it is customary to refer to the image impedance existing at a particular junction point as the *impedance level* at that point.

The image-impedance method of expressing the properties of a four-terminal network is extremely important because most four-terminal networks used in communication systems are operated under conditions that approach very closely image-impedance operation. This is particularly true of filters, equalizers, and impedance-matching networks.

Four-terminal Networks Operated on an Iterative-impedance Basis

A network operated on an iterative-impedance basis requires that the load impedance be such that the input impedance of the network with the load connected is equal to the load impedance. At the same time, the impedance observed by looking into the network from the output terminals toward the generator with the internal impedance of the generator connected across the input terminals of the network must equal the generator impedance. These two impedances can be designated as Z_{k_1} and Z_{k_2}, respectively, and are properties of the network. The third property necessary to specify completely the network characteristics is then taken as the iterative transfer constant P, which is defined by the equation

$$\frac{I_2}{I_1} = \epsilon^{-P} \tag{22}$$

where I_1 and I_2 are the input and output currents of the network, respectively, when operated under iterative conditions. Iterative-impedance action becomes of importance in handling problems involving L and ladder types of attenuators.

Impedance Matching

Two impedances are said to be matched when they have the same magnitude and the same phase angle. Thus in Fig. 10(a), the load impedance Z_L is said to be matched to the generator impedance Z_s if the load

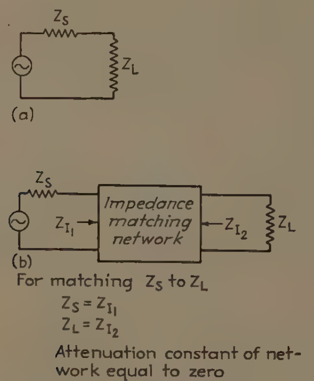


Fig. 10—Impedance matching of generator and load with the aid of a network.

impedance equals the generator impedance. If the load and generator impedances are not matched, it is possible to obtain matching by inserting between generator and load, as shown in Fig. 10(b), a network having $Z_{I_1}=Z_s$, $Z_{I_2}=Z_L$, and zero attenuation (i.e., real part of image transfer constant θ must equal zero). The generator then sees a load impedance $Z_{I_1}=Z_s$, and the load receives its power from a source (the network output terminals) having an internal impedance $Z_{I_2}=Z_L$. Such a system has its impedances matched on an image basis at all junction points.

The ratio of load current that would be delivered by a particular generator to a particular load without matching, as indicated in Fig. 10(a), to the same current when the impedances are matched, as in Fig. 10(b), is designated by such terms as *mismatching factor*, *reflection factor*, or *transition factor*. The absolute value of this ratio gives the loss of load current that results when no means are provided to couple the generator to the load impedance on an image-impedance basis. The value of the mismatching factor depends only upon the ratio of load to generator impedance, and is

$$\frac{\text{load current without matching}}{\text{load current with matching}} = k = \frac{\sqrt{\dfrac{4Z_s}{Z_L}}}{1+\dfrac{Z_s}{Z_L}}. \qquad (23)$$

Expressed in decibels, the loss of load current resulting from mismatching is

$$\text{mismatching loss in db} = 20 \log_{10} (1/k) \qquad (24)$$

A chart giving loss from mismatching based on (24) is given in Fig. 11, where θ_s and θ_L are the phase angles of Z_s and Z_L, respectively.

When the generator impedance Z_s is a resistance, then image-impedance matching corresponds to the condition for which the power delivered by the generator to the load is maximum, and if the load impedance fails to match the generator impedance, a loss of load power results. However, if the generator impedance has a reactive component, then failure to match the load impedance to the generator impedance can, under certain conditions, result in an increase in the load current.[10] Under these conditions, the mismatching factor will be greater than unity.

Insertion Loss

When a generator is inserted between a sending-end impedance Z_s and a load impedance Z_L, as in Fig. 10(b) the ratio of current in the load impedance when the

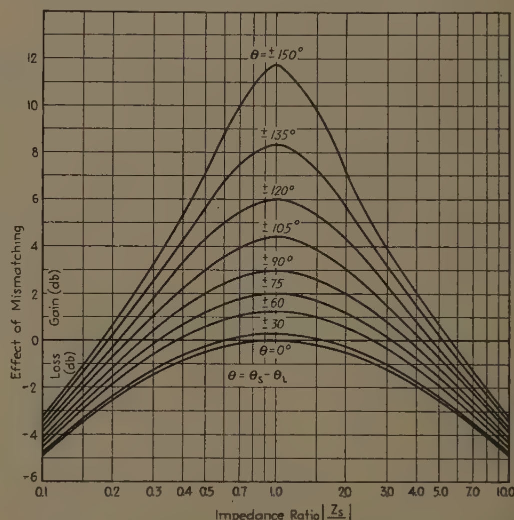


Fig. 11—Chart giving effect of mismatching generator and load impedance.

network is present to the load current in the absence of the network is termed the *insertion loss*, since this is the loss in output current resulting from the insertion of the network between generator and load.

The insertion loss can be conveniently expressed by the formula

[10] The condition for maximum possible transfer of energy to the load is realized when the resistance component of the load impedance is equal to the resistance component of the generator impedance and when at the same time the reactive component of the load impedance is equal in magnitude but opposite in sign to the reactive component of the generator impedance.

$$\text{insertion loss} = \frac{k_1k_2}{k}\sigma\epsilon^{-\theta} \tag{25}$$

where k_1 = mismatching factor of Z_s and Z_{I_1}
k_2 = mismatching factor of Z_{I_2} and Z_L
k = mismatching factor of Z_s and Z_L
θ = image transfer constant
σ = interaction factor

$$= \frac{1}{1-\left(\frac{Z_{I_2}-Z_L}{Z_{I_2}+Z_L}\right)\left(\frac{Z_{I_1}-Z_s}{Z_{I_1}+Z_s}\right)\epsilon^{-2\theta}}.$$

The interaction factor σ is a second-order effect representing a modification of the insertion loss that occurs when there is mismatching at *both* the input and output terminals of the network. The interaction factor takes into account the effect of a wave that is reflected from the load back to the generator through the network, and there reflected back to the load. The interaction factor becomes unity whenever at least one end of the network is matched on an image-impedance basis, or when the network attenuation is such that a wave that has traveled through the network twice will have been reduced to negligible amplitude.

Under practical conditions, the interaction factor in networks intended to be operated on an image-impedance basis is at most only a few decibels.

VII. Fundamental Types of Four-terminal Networks

π and T Networks

The fact that any four-terminal network can have its properties represented, insofar as the terminals are concerned, by three independent constants means that such networks can be always represented by three properly chosen independent impedances arranged in the form of a T or π, as shown in Fig. 12.

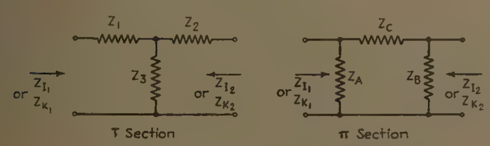


Fig. 12—General T and π networks.

The relationship between the impedances composing such networks, and the characteristics of the system expressed on an image-impedance basis, are

For T section (see Fig. 12):

$$\left.\begin{aligned}
Z_{I_1} &\quad \sqrt{(Z_1Z_2 + Z_1Z_3 + Z_2Z_3)\left(\frac{Z_1+Z_3}{Z_2+Z_3}\right)}\\
Z_{I_2} &= \sqrt{(Z_1Z_2 + Z_1Z_3 + Z_2Z_3)\left(\frac{Z_2+Z_3}{Z_1+Z_3}\right)}\\
\tanh\theta &= \sqrt{\frac{(Z_1Z_2 + Z_1Z_3 + Z_2Z_3)}{(Z_1+Z_3)(Z_2+Z_3)}}
\end{aligned}\right\}. \tag{26}$$

For π section (see Fig. 12):

$$\left.\begin{aligned}
Z_{I_1} &= Z_A\sqrt{\frac{(Z_B+Z_C)}{(Z_A+Z_C)}\frac{Z_C}{Z_A+Z_B+Z_C}}\\
Z_{I_2} &= Z_B\sqrt{\frac{(Z_A+Z_C)}{(Z_B+Z_C)}\frac{Z_C}{Z_A+Z_B+Z_C}}\\
\tanh\theta &= \sqrt{\frac{Z_C(Z_A+Z_B+Z_C)}{(Z_A+Z_C)(Z_B+Z_C)}}
\end{aligned}\right\}. \tag{27}$$

In the design of T and π networks, one normally knows the desired image impedances, and wishes to realize a particular transfer constant θ. The relations are then

For T section (see Fig. 12):

$$\left.\begin{aligned}
Z_3 &= \sqrt{Z_{I_1}Z_{I_2}\left(\frac{1}{\tanh^2\theta}-1\right)}\\
Z_2 &= \frac{Z_{I_2}}{\tanh\theta} - Z_3\\
Z_1 &= \frac{Z_{I_1}}{\tanh\theta} - Z_3
\end{aligned}\right\}. \tag{28}$$

For π section (see Fig. 12):

$$\left.\begin{aligned}
Z_C &= \sqrt{Z_{I_1}Z_{I_2}}\sinh\theta\\
Z_B &= \frac{1}{\frac{1}{Z_{I_2}\tanh\theta}-\frac{1}{Z_C}}\\
Z_A &= \frac{1}{\frac{1}{Z_{I_1}\tanh\theta}-\frac{1}{Z_C}}
\end{aligned}\right\}. \tag{29}$$

Examination of these equations shows that in the case of networks with reactive elements, the image impedances are either pure resistances or pure reactances. Furthermore, when the image impedances are resistive, the image transfer constant is a pure imaginary, while when the image impedances are reactive, the image transfer constant has a real component, and so attenuation is introduced. These relations are particularly important in the case of filters and impedance-matching networks.

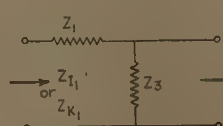


Fig. 13—General L network.

L Networks

An L network is shown in Fig. 13 and can be considered as a special case of a T or π network in which one of the impedance arms has become either zero or infinity.

The properties of an L network can be expressed in terms of image impedances and a transfer constant, as in the case of any four-terminal network. However, since there are only two impedance arms in an L network, a relationship must exist between the image impedances and image transfer constant such that if two of these are defined the third is likewise determined.

The formulas relating the image impedance and image transfer parameters of an L network with the impedance elements of the L are

$$\left.\begin{aligned} Z_1 &= \sqrt{Z_{I_1}(Z_{I_1} - Z_{I_2})} \\ Z_3 &= Z_{I_2}\sqrt{\frac{Z_{I_1}}{Z_{I_1} - Z_{I_2}}} \\ \tanh\theta &= \sqrt{\frac{Z_{I_1} - Z_{I_2}}{Z_{I_1}}} \end{aligned}\right\}. \tag{30}$$

The corresponding formulas for iterative impedance operation are

$$\left.\begin{aligned} Z_1 &= Z_{k_1}\left(\frac{\epsilon^P - 1}{\epsilon^P}\right) \\ Z_3 &= \frac{Z_{k_1}}{\epsilon^P - 1} \\ Z_{k1} &= \frac{Z_{k_1}}{\epsilon^P} \end{aligned}\right\} \tag{31}$$

where Z_{k_1} and Z_{k_2} are the iterative impedances of the two ends of the network, as shown in Fig. 13, and P is the iterative transfer constant, as defined by (22). L networks find their chief use in impedance-matching systems and in attenuators.

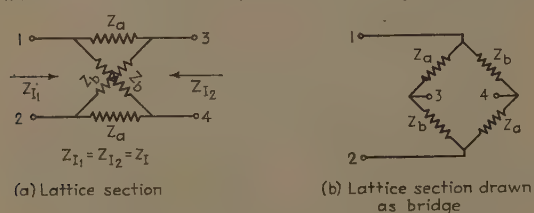


Fig. 14—General lattice network.

Lattice Sections

A lattice is a symmetrical balanced four-terminal network composed of two pairs of impedances, arranged as shown in Fig. 14. It will be noticed that the lattice is essentially a bridge in which the input is applied across one diagonal of the bridge, and the output is taken from the other diagonal.

The basic formulas of the lattice section in terms of image impedance and image transfer constant are

$$\begin{aligned} Z_I &= Z_{I_1} + Z_{I_2} = \sqrt{Z_a Z_b} \\ \tanh\left(\frac{\theta}{2}\right) &= \sqrt{\frac{Z_a}{Z_b}}. \end{aligned} \tag{32}$$

If one is given θ and Z_I, then

$$\begin{aligned} Z_a &= Z_I \tanh\left(\frac{\theta}{2}\right) \\ Z_a &= \frac{Z_I}{\tanh\left(\frac{\theta}{2}\right)}. \end{aligned} \tag{33}$$

It is possible to represent any symmetrical four-terminal network by a lattice having physically realizable impedance arms. In contrast, the π- or T-network equivalent of a complicated four-terminal network will sometimes require negative circuit elements in some of the arms, and hence be physically unrealizable. The image impedance of a lattice depends only upon the product of the two branch impedances, whereas the image transfer constant depends only upon the ratio of these impedances. It is therefore possible in the lattice to control the transmission characteristics entirely independently of the image-impedance behavior.

Lattice networks are used in filters and equalizers.

VIII. Reactive T, L, and π Networks for Matching Impedances[11–14]

T and π Reactive Networks

T and π networks having impedance arms composed of reactive elements are widely used for matching an antenna to a transmission line in order to give a nonresonant characteristic-impedance termination for the line. It is possible with such networks to transform any resistance load that may be offered by the antenna system to any other value of resistance that may be needed to give a characteristic-impedance load for the transmission line. At the same time, the phase shift introduced by the impedance-matching network can have any desired value.

On the assumption of an ideal network composed of reactive impedances with zero losses, the design relations represented by (28) and (29) can be written as

For T section:

$$\left.\begin{aligned} Z_1 &= -j\,\frac{R_1\cos\beta - \sqrt{R_1R_2}}{\sin\beta} \\ Z_2 &= -j\,\frac{R_2\cos\beta - \sqrt{R_1R_2}}{\sin\beta} \\ Z_3 &= -j\,\frac{\sqrt{R_1R_2}}{\sin\beta} \end{aligned}\right\} \tag{34}$$

[11] W. L. Everitt, "Output networks for radio-frequency power amplifiers," Proc. I.R.E., vol. 19, pp. 725–738; May, 1931.

[12] W. L. Everitt, "Coupling networks," *Communications*, vol. 18, p. 12; September, 1938; and p. 12; October, 1938.

[13] Carl G. Dietsch, "Terminating concentric lines," *Electronics*, vol. 9, p. 16; December, 1936.

[14] Ralph P. Glover, "R-f impedance matching networks," *Electronics*, vol. 9, p. 29; January, 1936.

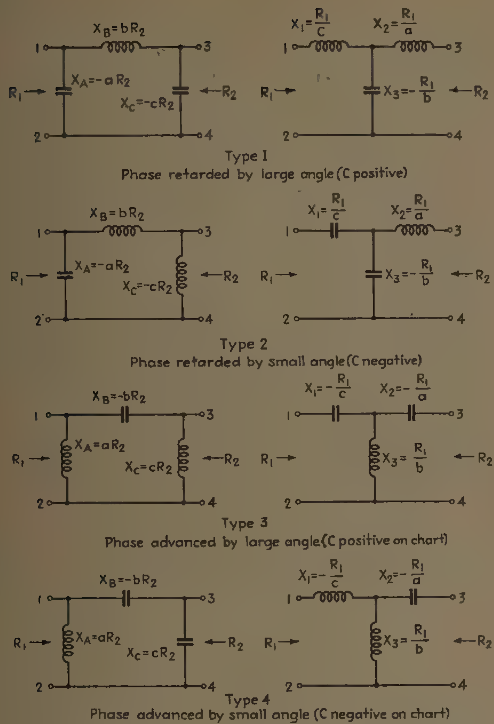


Fig. 15—Three-element reactive networks which may be used for impedance matching.

For π section:

$$\left.\begin{aligned} Z_A &= j\frac{R_1R_2 \sin\beta}{R_2\cos\beta - \sqrt{R_1R_2}} \\ Z_B &= j\frac{R_1R_2 \sin\beta}{R_1\cos\beta - \sqrt{R_1R_2}} \\ Z_C &= j\sqrt{R_1R_2}\sin\beta \end{aligned}\right\} \quad (35)$$

where R_1 and R_2 are the two image impedances and β is the angle by which the phase at the output terminals lags behind the phase at the input terminals. Design charts derived from these equations are given in Figs. 16, 17, and 18, and together with Fig. 15 will give the reactances required. The charts can be used for negative as well as positive values of β. For negative values of β, the magnitudes of the constants a, b, and c are the

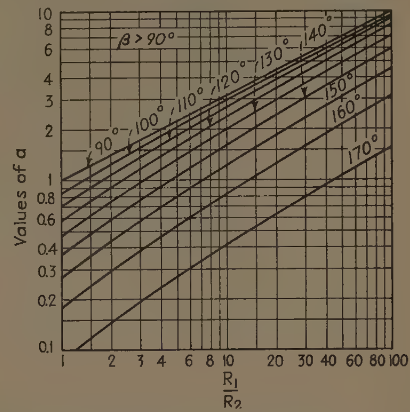


Fig. 17—Design charts, giving values of b for different values of β for use in networks of Fig. 15.

same as for positive, but the signs used in front of the constants are reversed, as shown in Type 3 and 4 networks of Fig. 15. In the design curves of Figs. 16 to 18, it has been assumed that R_1/R_2 is greater than unity. This is no restriction, since terminals 1–2 can be placed at either the generator or the load end of the network, according to whichever must match the higher resistance.

L Reactive Networks

An L network composed of reactive elements is able

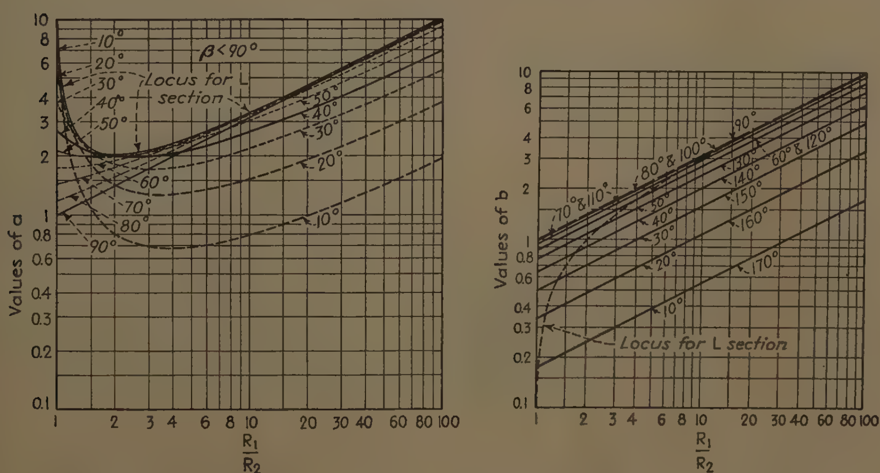


Fig. 16—Design charts, giving values of a for different values of β for use in networks of Fig. 15.

to transform a given resistance to make it look like any other resistance by making the image impedance of the network at the terminals facing the load equal the load resistance and the image impedance at the other terminal equal the desired resistance. The phase shift introduced by an L section is determined by the ratio of image impedances, and cannot be specified independently, because the L section has only two impedance arms. The L section can be considered as a special case of Fig. 15, for which constant c has the value $c=\infty$, leading to sections as in Fig. 19. The loci corresponding to this condition are shown on Figs. 16 and 17. With an L section, the maximum phase shift obtainable is ± 90 degrees.

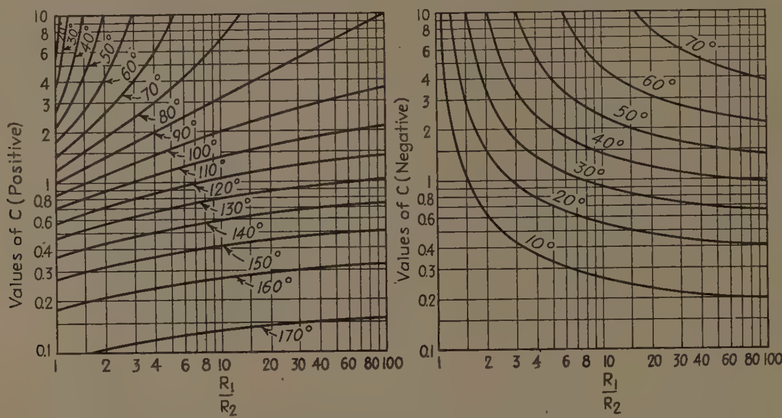


Fig. 18—Design charts, giving values of c for different values of β for use in networks of Fig. 15.

Dissipation of Power in Reactive Networks

In practical networks, the condensers have neglibible loss, but the resistance of the inductances is not entirely negligible. On the assumption that the currents in the various network branches are not appreciably affected by the dissipation of the inductive elements and that the various inductive elements have a ratio of reactance to resistance (i.e., Q) that is the same for all inductances, one can write[11,12]

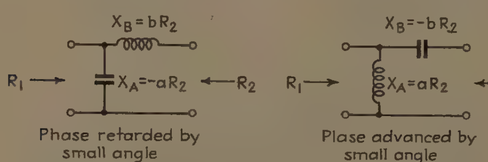


Fig. 19—Section of L type that is obtained when c in Fig. 15 becomes infinite.

$$\frac{\text{power lost in network}}{\text{power delivered to network}}=\frac{\delta}{Q} \tag{36}$$

where δ is a constant given by either Fig. 20 or 21, according to the type of section involved. These curves apply for L as well as T and π networks, the locus for the L network being dotted on the curves.

A study of Figs. 20 and 21 shows that the efficiency of the network is implicitly determined by the impedance transformation ratio and the phase shift. There is no choice between T and π networks and between networks that advance and retard the phase, insofar as efficiency is concerned. The loss increases with increasing transformation ratio, and tends to be large when the phase shift of the network is either very small or very large. Finally, it will be noted that for a given transformation ratio, the L section has a lower loss than either the T or π section. In cases involving very high transformation ratios, or phase shifts that are either very small or approach 180 degrees, an increase of efficiency can be obtained by dividing the total impedance transformation and total phase shift among two or more networks connected in tandem.

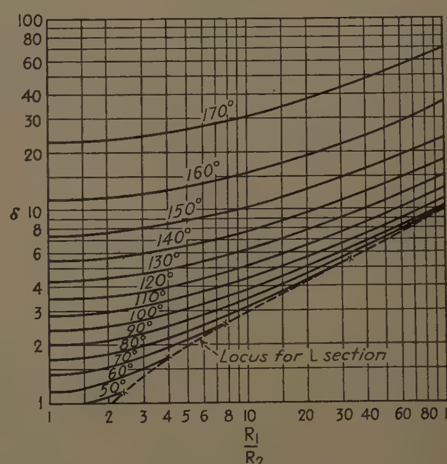


Fig. 20—Values of δ for use in (36), applicable for Type 1 and Type 3 networks of Fig. 15. The angles are values of β.

Harmonic Reduction

Coupling networks that call for an inductance in the series arm or a capacitance in a shunt arm, or both, can be readily arranged to provide discrimination against harmonics. The discrimination against a particular harmonic can be made particularly great when the series inductive reactance is supplied by a parallel circuit, as shown in Fig. 22 (b), resonant at a frequency to be suppressed and so proportioned as to give the required inductive reactance at the frequency to be transmitted. An equivalent result is also obtainable when the shunt capacitive reactance is supplied by a suitably designed series circuit, as in Fig. 22(c), that is resonant at the frequency to be suppressed.

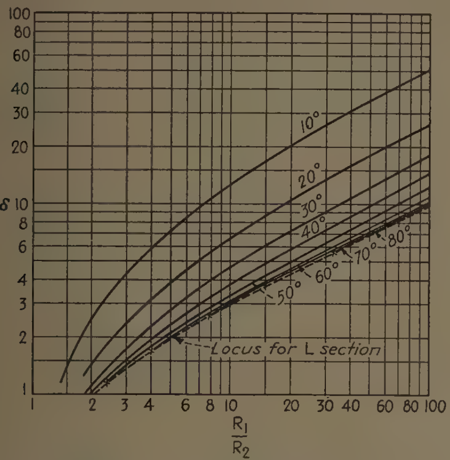


Fig. 21—Values of δ for use in (36) applicable for Type 2 and Type 4 networks of Fig. 15. The angles are values of β.

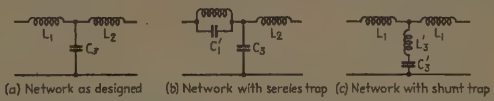


Fig. 22—Diagram illustrating how a series inductance and a shunt capacitance of an impedance-matching network may be replaced by parallel- and series-resonant circuits respectively, to increase discrimination against an undesired harmonic.

The design formulas are as follows:

For Fig. 22(b):

$$L_1' = L_1(1 - \gamma^2) \tag{37a}$$

For Fig. 22(c):

$$C_3' = C_3(1 - \gamma^2) \tag{37b}$$

where γ is the ratio of the frequency to be transmitted to the frequency to be suppressed. C_1' in Fig. 22(b) and L_3' in Fig. 22(c) are assigned values that will make the resonance occur at the frequency to be suppressed.

(To be continued)

Address of Retiring President*

ARTHUR VAN DYCK†, FELLOW, I.R.E.

THE YEAR 1942 was extraordinary in the record of Institute history. It was the thirtieth anniversary of the founding of the Institute. It was the first year of World War II. It marked the end of one era in radio, and the beginning of another. The year has been packed with events of great significance to radio, and consequently to The Institute of Radio Engineers. There have been not only technical developments, but there have occurred the beginnings of changes and movements which will have vast effect upon radio science and industry after the war. A complete review of the year is impossible in such a report as this, and I can hope only to select a few high lights. Even a selection of items is difficult, and I must confess to some bewilderment in rating and choosing subjects on the basis of importance.

One subject of certain interest and importance is that of the Institute itself, and its progress during the year, in membership, organization, and accomplishment. Full reports thereon will be published in the PROCEEDINGS, and it is not necessary here to report in detail. In connection with membership, our roster now totals 8775 members, which is a growth during the year of approximately 25 per cent. This has been a natural and unforced increase, without membership drives and pressure solicitation. I hope that during 1943, it will be appropriate and possible to increase membership by a much greater percentage, from the ranks of the thousands of newcomers to radio in war services, and I recommend active membership campaigns, not only by the regular committees, but by each Section and by every individual member.

It should be recorded that the quality of the PROCEEDINGS has been maintained. In spite of extra wartime difficulties, particularly those of papers securement and of censorship, the PROCEEDINGS have been issued regularly, on time, with papers having high interest and value. This record has been achieved by much extra work on the part of the Editors, particularly their Chairman, Dr. Alfred N. Goldsmith, and by the efforts of the Papers Procurement Committee,

* Decimal classification: R009. Original manuscript received by the Institute, January 26, 1943. Presented, Winter Conference, New York, N. Y., January 28, 1943.
† RCA Laboratories, New York, N. Y.

headed by Mr. Dorman Israel. I wish to express the thanks we all feel to them for their unselfish devotion to this work which means so much to the profession. There are good prospects that they have planned and built, and are planning and building, so well, that 1943 will see continuance of high quality in the PROCEEDINGS in spite of the difficulties introduced by war conditions.

A number of changes in office personnel were brought about during the year by war conditions. Early in the year Mr. John D. Crawford resigned to enter war engineering work. Also early in the year, the Secretary, Mr. Harold Westman, began extra work in standardization of radio material, for the War Production Board and the American Standards Association. This work developed in magnitude and importance until toward the end of the year his full time was required by the American Standards Association. Consequently, he resigned the Secretaryship of the Institute, and while we are sorry to lose him, we can view his loss to us as a distinct gain to the war effort, because it is certain that he is uniquely qualified for the difficult work of radio standardization. Already, valuable results have been accomplished under his leadership, and much more will follow.

In further reference to office operations, I must report that the Directors, particularly those who have served on the Executive Committee, have worked hard and long during this difficult year. The loss of the Assistant Secretary and the part-time service of the Secretary placed extra work upon the Executive Committee. In spite of the fact that each member was already heavily loaded by war work, each one responded cheerfully to every call to more work for the Institute. I would like to mention especially the services of Mr. Haraden Pratt, who has given unstintingly of his time and energy. His quiet wisdom, good judgment, and unselfish willingness to help in anything asked of him, have meant much to the welfare of the Institute, and to me personally in my job, throughout the year.

The financial condition of the Institute continues to be sound, as a result of prudent, good engineering management. The Institute has weathered depression years and one year of war without financial difficulty. I would like to point out, however, that the income of the Institute is not such as will permit engaging in activities for the welfare and advancement of the profession, such as may become advisable in the postwar period. Consequently I recommend that study be given during the coming year to ways and means of increasing income in order that the Institute may have the ability to enlarge its activities in keeping with the enlarged importance of radio and its engineering profession.

During the year, the Board of Directors has given much study to factors basic to the preparation of the Institute for the difficult years ahead. The expansion of radio technique, the acceleration thereof by war necessities, and the war itself, have brought about changes and new conditions, which must be recognized and prepared for, if the Institute is to progress or even to retain the eminence in its field which it has held heretofore. Not to progress is to retrogress, and the next year or two will be a critical time, in which the long-term future of the Institute will be largely influenced, if not completely determined.

As a result of its study, the Board has evolved, and is submitting to the membership, proposals to change the existing status of membership requirements in allied communication-and-electronic fields to one of equality, and to change the membership-grade structure by the addition of another grade, intermediate between the present Member and Associate grades. It is unnecessary to discuss these proposals at this time, but I do recommend them to you for approval, as having potentialities of benefit to the Institute under future conditions.

During the year, the organization of a New York Section was completed, and it has begun operations with vigor and vitality. It is believed that this will not only relieve the National Office of the work connected with New York meetings, but will give new force and incentive to this area, which includes a high percentage of the membership.

The year witnessed remarkable revolution in the radio industry, in which its engineers were called upon for heroic performances. Production of peacetime civilian-radio apparatus was completely stopped, and the manufacturing capacity of the industry was converted to military production within a fraction of the year. It was not easy to make such a change, for the designs and the manufacturing processes and tolerances are quite different for devices in the home and devices in fighting machines. But it was done, and now 350,000 persons are engaged in the production of precision radio for our fighting forces.

It should be mentioned also that the broadcasting side of the industry has kept pace with requirements. Not only has it maintained its home programs in quantity and quality in spite of personnel and materials difficulties, but it has greatly increased its transoceanic services, with their important uses of home contact to our men overseas, and the spreading of information on the American way to all peoples everywhere.

There are several matters to which I would like to call attention, although they cannot be classified as accomplishments of 1942. They appeared in 1942, however, and their treatment is in the future. I believe that they should be mentioned at this time, in order that we may keep before us the agenda which we shall have to meet.

First there is the radio personnel problem. That has a current aspect and a postwar one. Its present importance involves training, selective service, and allocation. To the Institute, it presents an opportunity

both to serve the country, and to benefit itself by increased membership. Its postwar importance will result from the condition that the number of workers in radio at the end of the war will be many times that in it before the war. It is doubtful that commercial radio will be able to utilize immediately all those who will wish to enter it, and there will be problems of selection and allocation of personnel. If the Institute can be prepared to assist in meeting these problems, it will be highly desirable to do so.

After the war, there will be a sudden and huge release of technical papers for publication. Many developments have occurred in each branch of radio technique, but nearly all have been shrouded in war secrecy. When the need for secrecy is removed, literally hundreds of papers will be made available for publication. Inasmuch as the PROCEEDINGS of the Institute has always been a foremost publication in the field of radio, radionics, and electronics, it is advisable that preparations be made to handle this flood of papers efficiently and well.

Next, and of great importance, is postwar planning of certain technical matters, largely of the nature of standardization. For example, television should have review and speedy settlement, because that branch of radionics will be one of the few new industries capable of immediate utilization on a large scale in the conversion of industry from war to peace. The Chairman of the Board of War Communications has already sugfested that this be done, and I believe it is the duty of the Institute, as the professional radio engineering body of the country, to take leadership in such technical problems of the industry. Incidentally, I hope that we may achieve the benefits of international standardization, as well, so that the introduction of international television, after a few years, will not be impeded by unnecessary differences.

It may seem inadvisable, or at least premature, to consider postwar planning before we have won the war. Of course we must not let such planning interfere with the prosecution and successful conclusion of the war with maximum speed, but it is possible even at this time to do some things without penalty to the war effort. The radio industry did an excellent job of converting from peace to a war footing. That resulted from preparations made a year or more in advance. Similar foresight and planning will be necessary to assure that the conversion of industry to peace shall be as efficient and orderly as possible, with a minimum of unemployment, social disturbance, and economic dislocation.

And now engineering and engineers, science and scientists, have come to a crucial point in the development of civilization. Scientists and engineers have brought about progress which has culminated not in greater freedom, happiness, and security, but in the most destructive conflict the world has seen, one so destructive that it is destroying not only human lives, but whole empires, political castes, and social systems. Having made possible such efficient destruction, engineers and scientists should decide what they are going to do next. Of course, they may let the decision go by default, and let nature take its course as they have done in the past. On the other hand, they have the opportunity now to decide to do something in the future which will encourage happier results and safer progress in the development of civilization. Something must be done in the future, which will give saner judgment and more scientific control of the forces unleashed by science, if we are to progress.

Many will say that it is presumptuous and egotistical for engineers to talk or think in this way. To those I would say, listen to the words of many leaders of thought today, who are not engineers. For example, first, Dorothy Thompson—"What stands out in ever more glaring relief is the gap between scientific accomplishments and imagination regarding the uses to which they are put. The greatest star in the crown of twentieth-century capitalism is its creation of our industry. The greatest blot on it is its recurrent failure to put the public interest, necessity, and welfare in the foreground of its thinking and planning. It has technological foresight out of all proportion to its social foresight."

Next, from an article in *Fortune* magazine—"Technology is simply the application of the sum of man's knowledge of the physical world to the task of getting a job done with maximum results and a minimum of error. It is science in action. It is now a decisive element in the rise and fall of civilizations. The points at which technical decisions are made today are the most critical spots in the world."

Next, Professor Bridgman of Harvard—"It seems to me that scientists are curiously obtuse as to the social conditions which make possible their existence as a class. It is by no means a certainty that society will so evolve that the individual will be allowed to engage in independent intellectual activity. Society may well come to feel that the scientist has not enough more to give it in the way of material benefits, to justify keeping him. We must do certain things if we are to play our part in molding a public opinion which will create the society of our vision."

Next, a *New York Times* editorial— ". . . American scientists and engineers have not given nearly so much thought to the scientific and technological implications of the Atlantic Charter as their British colleagues. We have some fine American discussions of postwar economics, but no clear program to indicate the part that science and technology must play in removing the economic causes of war.

"The turning point in British thinking came in 1932 when the International Congress for the History of Science was held in London . . . . Militant English scientists began to discuss the frustration of science and to urge laboratory workers to think of themselves

as citizens of the world whose plain duty it was to make their influence felt.

"Out of this agitation came the decision of the British Association for the Advancement of Science to consecrate a whole division to the planned, global application of science to social progress. The American Association for the Advancement of Science followed with a decision to co-operate with the British association in studying what we now call the 'impact of science on society.'

"There is every reason why science should be organized internationally to implement the fourth article of the Atlantic Charter. Science has always been international. Its votaries accept one another for what they know and for what they have achieved. At international scientific meetings no questions are asked about race, nationality, or creed. Each man gives freely for the benefit of the world.

"Scientists still stand apart in their internationalism because their objective mode of thinking is not yet the common possession of mankind. If ever we succeed in implementing the fourth article of the Atlantic Charter by granting not only equal access to raw materials but equal access to the technical knowledge by which raw materials are exploited, we shall go far toward removing the intellectual barriers that separate the laboratory from the common mind.

"Out of the British conference on international resources of last July two axioms emerged. The first is this: We can no longer afford to leave the use of world resources and of human ingenuity and skill to blind economic and national forces. The alternative to international co-operation is the old chaos.

"Planning inevitably leads to the second axiom, which is this: An International Resources Office must be established, an office which will concern itself not only with postwar relief and postwar reconstruction but with the perpetual research and the application of scientific knowledge so that not only actual raw materials (petroleum, coal, metals, plant and forest products) but the techniques whereby these raw materials can be best utilized for the general good of the world may become accessible."

In speaking of these things, I hope not to be thought pedantic, but I do consider it my duty as your President for a term, to give you my observation and opinion that these matters have assumed importance greater than that of the technical things to which we have heretofore confined ourselves, that we are by training and experience better able to cope with them than others not so trained, and lastly, that if we do not exhibit interest and activity in them, we shall find ourselves in a world of chaos. Isolationism is now impossible, for scientists as well as for nations.

Perhaps we can see a good lesson in the current performance of Soviet Russia. Many have wondered at the remarkable way in which that country has turned back the mighty German military machine. Perhaps the secret lies in certain facts revealed in recent studies by the New School for Social Research, which show that the long experimentation of the Soviet in government and industrial methods finally developed from 1936 to 1938 into a plan wherein economic control and industrial management were unified in the hands of young production engineers. This change has gone so far that nearly one third of the government offices are held by engineers. These men are not only better fitted technically to handle a technical economy, but are trained to decide questions on a basis of facts rather than of political expediency. The spirit of Russia was implemented by technically sound administration, and was thereby made effective.

As my retiring message, I submit to you the proposition that the time has come for us to think of larger responsibilities, to seek them, and to accept them. Only by so doing can we do our full duty toward making this world safe for civilization.

I desire to thank the membership for the honor and privilege of serving as President. It has been a pleasure to serve, even though it has been a war year, and not a peaceful, placid one. It has been a pleasure because it has meant close association with the Directors, a group of men whom I wish you all could know as I have come to know them—sincerely interested in the welfare of the Institute, and unstinting in time and effort to promote that welfare. They remain, among them your new President, and in his and their hands during this next year, I shall know that the Institute is going forward in the bigger and better future which awaits it.

Institute News and Radio Notes

Postwar Horizons

Important contributions to be made by the radio-and-electronic field toward postwar economic recovery and renewed prosperity were envisioned by David Sarnoff, President of the Radio Corporation of America and former Secretary of The Institute of Radio Engineers, in an address delivered before the Chamber of Commerce of the State of New York on February 4, 1943. He said in part:

"Our hope for a future world economy of abundance is founded upon much more than prewar standards of prosperity. It is based upon the promise of industrial science. The old frontiers of the world were frontiers of geography. The new frontiers are those of science. The covered wagon of the present day is the research laboratory.

"Progress in the field of radio and electronics has advanced on the same broad front with progress in other fields of science and industry. It is radio which has made possible a war of speed and mobility on land, at sea and in the air. Radio-electronic sentinels stand watch on shipboard and along the coast. The United States now has fighting forces stationed at more than sixty strategic locations on the world map. Its Navy operates on the Seven Seas. Without instant, reliable radio communication it would be impossible for these widespread forces to function as a unified war machine."

Referring to the radio tube as the heart of every radio instrument, he pointed out that science in putting electrons to work in the tubes has greatly extended the usefulness of electronics in industry as well as in communications. He added further:

"We began learning how to control the elusive electrons in vacuum tubes, forty years ago. The versatility of these tubes, and of the devices built around them, is amazing. They can be made to respond to light, to all shades of color, to smoke, to the faintest noise. In terms of results, we can say that they are able to hear, see, feel, taste, remember, calculate, and even talk. They bring increased speed, accuracy, and safety to a wide variety of industrial operations."

Calling attention to the latest radio-electronic developments, Mr. Sarnoff appraised the electron microscope as an outstanding achievement. Capable of magnifying 100,000 diameters, it has opened new worlds of knowledge in biology, bacteriology, medicine, physics, chemistry, plastics, textiles and, other fields of research. In this regard, he stated:

"In most industries the emphasis is on bigness. Radio science is built on minuteness. An electron is a tiny fraction of an atom."

And to illustrate this point he called attention to the fact that the electron microscope made it possible to photograph the influenza virus for the first time. It should not be forgotten that a single invisible germ sometimes carries more power of destruction than a 2000-pound bomb.

Turning to another new field of radio, his analyses showed that radiothermics, the application of heat generated by high-frequency radio waves, is finding new and widespread use in speeding and improving industrial processes. For example, a laminated airplane propeller is processed in minutes compared with hours required by older methods. Similarly, radio heat may be employed to bond rubber to wood or plastic surfaces, to dry textiles and purify food products. Accordingly he foresaw radiothermics as a significant factor in postwar industry.

Describing television as "the most spectacular development in the field of communication," which may be looked forward to in the post-war period, he predicted that when the war is over, television will advance as a new service of public information and entertainment.

"We expect to have intercity networks of stations as we have them in sound broadcasting. Eventually they will become nation-wide. We look forward to television programs in theaters as well as in homes. Thanks to war research, these television pictures will be technically much better than they were before the war. It is gratifying to those who labored many years to bring television out of the laboratory to know the experience gained from television research is proving of vital importance in the war."

Board of Directors

The regular meeting of the Board of Directors, held on February 3, 1943, was attended by L. P. Wheeler, president; F. S. Barton, vice president; S. L. Bailey, E. F. Carter, I. S. Coggeshall, H. T. Friis, Alfred N. Goldsmith, editor; G. E. Gustafson, R. A. Heising, treasurer; Haraden Pratt, secretary; H. M. Turner, A. F. Van Dyck, H. A. Wheeler, W. C. White, and W. B. Cowilich, assistant secretary.

Approval was granted to the following applications: for transfer to Member from H. E. Dinger, Beverly Dudley, T. M. Ferrill, Jr., J. B. Knight, Clarence Radius, A. G. Sparling, C. J. Terrey, and R. G. Zender; for admission to Member from J. A. Cose, D. M. Davis, J. E. Innes-Crump, G. F. Maedel, O. S. Meixell, A. C. Omberg, Charles Sheer, and W. D. Wenger; and 110 for admission to Associate, 120 for Student, and 5 for Junior.

President Wheeler briefly reported on recent Executive Committee actions involving temporary changes in the Code of Administrative Practice, the appointment of three members to be responsible for the Institute's committee activities, and early conference with chairmen of existing Technical Committees relative to personnel of their committees for the next term, beginning May 1, 1943.

The editorial, "Wartime Service," by President Wheeler, was approved for publication in the March, 1943, issue of the PROCEEDINGS.

The bank resolutions relative to the "General Account" and to that "Special Account" requiring only the secretary's signature for withdrawal, were authorized.

The bank resolution regarding a second "Special Account" for office use and providing for withdrawal only on the signature of the assistant secretary, was rejected.

In place of the latter "Special Account," an account to be known as the "Office Account" was established for purposes of office administration. The bank resolution on this account was also approved.

The following resolution concerning the readmission of former members was voted:

> "Resolved, that the Board of Directors readmit to the grade of membership previously held (or in the Associate grade if formerly a Junior and now over 21 years of age) those former members (a) whose memberships terminated before or during 1942 and who pay either current dues or all dues in arrears or (b) whose memberships terminate on March 31, 1943 and who pay dues for 1943 at a later date during 1943. The payment of a new entrance fee, if such would normally be required, is waived. Associates, who formerly had the privilege of voting, will be readmitted as nonvoting Associates."

Reports were given on Executive Committee actions relative to PROCEEDINGS advertising and to the charge for the completed audit of the Institute's 1942 financial records.

The auditor's report for the fiscal year ending December 31, 1942 was accepted.

It was agreed to make certain changes in office practice. Consideration was given to the matter of changing the date of the first mailing of membership-dues bills, which is now December first as specified in the Constitution, in view of the year-end congestion of other work in the office.

The appointment of a committee to investigate the Institute's investments was delegated to President Wheeler and Treasurer Heising.

President Wheeler announced the two actions of the Executive Committee, one with reference to the proposals of B. J. Thompson, chairman of the Membership Committee, for conducting the activities of that group this year, and the other with reference to holding PROCEEDINGS issues in reserve for members in the armed services.

Two letters received in reply to the Institute's resolution on deferment of engineering students were read. They were from the Office of Emergency Management of the War Manpower Commission, and the Signal Corps Manpower Division of the Selective Service System, both at Washington.

(Continued on page 182)

CANDID CAMERA COMPLETES THE RECORD

Dr. Southworth addresses the joint I.R.E.-A.I.E.E. meeting at New York

Dr. Wilson receives the Medal of Honor from President Wheeler

Vice President Barton (right) converses with Past-President Van Dyck

OF THE WINTER CONFERENCES OF 1943

At Boston, Dr. DuBridge receives Fellow diploma from Professor Oberg (center). (Others, left to right, Messrs. Henry, McElroy, and Eastman and Commander Meader.)

Chairman Fly of the F.C.C. Addresses Washington Section and Nationwide I.R.E. audience. (Seated, left to right, General Stoner, Section Chairman Hunt, Admiral Ranneft, Captain Wilson, and Mr. McIntosh of W.P.B.)

(*Continued from page 179*)

The subject of more extensive promotion of the interests and welfare of the Institute and its membership was discussed.

The letters from C. L. Fortescue, president, and R. L. Smith-Rose, both of the Institution of Electrical Engineers, London, suggesting an exchange of that society's facilities with those of the Institute, were read by President Wheeler. This suggestion was accepted wholeheartedly.

A memorandum on the Institute's foreign-membership policy, previously prepared, was discussed.

President Wheeler called attention to the splendid co-operation received from the American Institute of Electrical Engineers during our Winter Conference at New York City, and to his recent letter to H. S. Osborne, their president, expressing the Institute's appreciation.

President Wheeler stated that the Institute had sent flowers to the funeral of the late Dr. Nikola Tesla.

Mr. Coggeshall, chairman of the Winter Conference Committee, reported on the conference held on January 28, 1943. Satisfaction with the success of that conference was expressed and Mr. Coggeshall was given a vote of gratitude.

The proposed Constitutional Amendments, which in due course will be submitted to the membership as required by the Constitution, and the resolution relative to their submission to legal counsel, are given below:

"Resolved: that the Board of Directors does hereby propose the following amendments to the Institute Constitution and does hereby instruct the Secretary to submit these proposed amendments to legal counsel for his opinion as to their accordance with the laws under which the Institute is organized:—"

Proposed Revision of Article II, Sections 1 through 8 and Section 10; Article III, Sections 4 and 7; Article IV, Section 1; Article V, Section 1; Article VI, Sections 4, 5, and 6; Article VII, Heading and Section 2 of the Constitution

Article II

MEMBERSHIP

Sec. 1—The membership of the Institute shall consist of:

a. Fellows, who shall be entitled to all rights and privileges of the institute.

b. Senior Members, who shall be entitled to all rights and privileges of the Institute except the right to hold the offices of President and Vice President.

c. Members, who shall be entitled to all rights and privileges of the Institute except the right to hold the offices of President, Vice President, and Director.

d. Associates, who shall be entitled to all rights and privileges of the Institute except the right to hold the offices of President, Vice President, and Director, and the chairmanships of standing Committees and of Sections, and the right to vote. However, Associates of record on March 1, 1939, shall have the right to vote so long as a continuous membership since that date is maintained. Furthermore, the restriction on holding chairmanships of standing committees and of Sections shall not be operative until January 1, 1945.

e. Students, who may participate in meetings, receive the Proceedings, and wear the badge of the Institute, but who shall have no other rights or privileges.

Sec. 2—Fellow: For admission or transfer to the grade of Fellow, a candidate shall be at least thirty-two years of age and shall be either:

a. An engineer or scientist in radio or allied fields. As such he shall have attained distinction in his profession and shall be eminently qualified to take responsible charge of important radio or allied work. He shall have been in the active practice of his profession for at least ten years, and shall have had responsible charge of important radio or allied work for at least three years.

b. A professor of engineering or of physical science. As such he shall have attained special distinction as an expounder of the principles of engineering or of science in radio or allied fields. He shall have had at least ten years' experience as a teacher of electrical or physical subjects and shall have had responsible charge, for at least three years, of a course in radio or allied subjects in a school of recognized standing.

c. A person who has done notable original work contributing to the advancement of engineering or science in radio or allied fields which has given him a recognized standing at least equivalent to that required for Fellow under paragraph "a" or "b."

d. A person regularly engaged in radio or allied work for at least ten years, who, by invention or contribution to the advancement of engineering or science in radio or allied fields, or to the technical literature thereof, has attained a standing at least equivalent to that required for Fellow under paragraph "a" or "b."

Sec. 3—Senior Member: For admission or transfer to the grade of Senior Member, a candidate shall be at least twenty-six years of age and shall be either:

a. An engineer or scientist in radio or allied fields. As such he shall have performed and taken responsibility for important engineering or scientific work in these fields and shall have been in the active practice of his profession for at least six years.

b. A teacher of radio or allied subjects for at least six years in a school of recognized standing. He shall have been in responsible charge of a major course in such fields.

c. A person regularly employed in radio or allied work for at least six years, who by invention or by contribution to the advancement of engineering or science in radio or allied fields, or to the technical literature thereof, has attained a standing equivalent to that required for Senior Member under paragraph "a."

d. An executive who, for at least six years, has had under his direction important engineering or research work in radio or allied fields and who is qualified for direct supervision of the technical or scientific features of such activities.

Sec. 4—Member: For admission or transfer to the grade of Member, a candidate shall be at least twenty-four years of age and shall be either:

a. An engineer or scientist in radio or allied fields. As such he shall have demonstrated competence in engineering or scientific work of professional character and shall have been in active practice of his profession for at least three years.

b. A teacher of radio or allied subjects for at least three years.

Sec. 5—Associate: For admission or transfer to the grade of Associate, a candidtate shall be at least eighteen years of age and shall be interested in the theory or practice of radio engineering or of the allied arts and sciences.

Sec. 6—Student: For admission to the grade of Student, a candidate shall be devoting a major portion of his time as a registered student in a regular course of study in engineering or science in a school of recognized standing. Membership in this grade shall not extend more than one and one-half years beyond the termination of his student status as described above.

Sec. 7—The term "allied" as used in this Constitution and Bylaws refers to electrical communication, electronics, and such other technical fields as are directly contributory to or derived from radio.

Sec. 8—The expression "school of recognized standing" includes only schools of college grade providing an engineering or scientific curriculum of not less than four years and granting degrees, and such other schools as may be so designated by the Board of Directors.

Sec. 9—Graduation from a radio or electrical course of a school of recognized standing shall be accepted as equivalent to one year's experience in radio or allied fields. Full-time graduate work, or part-time graduate work with teaching, in radio or allied courses in a school of recognized standing, shall be accepted as equivalent to professional experience, for a period not exceeding two years.

Sec. 10—(old Sec. 9 without change).

Sec. 11—The terms "member" and "membership" when printed without an initial capital where used in this Constitution and Bylaws include all grades.

Sec. 12—(old Sec. 11 without change).

Article III

ADMISSIONS, TRANSFERS, AND EXPULSIONS

Sec. 4—The admission fee and dues are payable on notification of election and if not received within six months from notification, the election shall be void.

Sec. 7—When a member's dues become three months in arrears his membership shall terminate. Subject to the approval of the Board of Directors, membership may be resumed on payment of all dues in arrears or on payment

of a new entrance fee and current dues.

For purposes of Article II, Section 1-d, resumption of membership as provided under Sections 6 and 7 of this Article shall not constitute maintenance of continuous membership.

ARTICLE IV

ENTRANCE FEES AND DUES

Sec. 1—The entrance fees, transfer fees, and annual dues shall be as follows:

ENTRANCE FEES

FELLOW	$5.00
SENIOR MEMBER	5.00
MEMBER	3.00
ASSOCIATE	3.00
STUDENT	NONE

The transfer fee from one grade of membership to another shall be the difference between the corresponding entrance fees except that there shall be no fee when transferring immediately from Student to Associate membership.

ANNUAL DUES

FELLOW	$10.00
SENIOR MEMBER	10.00
MEMBER	6.00
ASSOCIATE	6.00
STUDENT	3.00

ARTICLE V

OFFICERS

Sec. 1—The governing body of the Institute shall be the Board of Directors and shall consist of the President, Vice President, Secretary, Treasurer, Editor, nine elected Directors, five appointed Directors, and the two most recent past Presidents.

ARTICLE VI

MANAGEMENT

Transpose Sections 4 and 5

Sec. 6—All funds received by the Institute shall be deposited in an account requiring the signatures of at least two of the following for withdrawal: President, Vice President, Secretary, Treasurer, and Editor. Funds from . . . etc.

ARTICLE VII

NOMINATION AND ELECTION OF PRESIDENT, VICE PRESIDENT, AND THREE DIRECTORS, AND APPOINTMENT OF SECRETARY, TREASURER, EDITOR, AND FIVE DIRECTORS

Sec. 2—The Secretary, Treasurer, and Editor, shall be appointed by the Board of Directors at its annual meeting to serve until the next annual meeting.

Executive Committee

The Executive Committee met on January 29, 1943, and those present were L. P. Wheeler, chairman; Alfred N. Goldsmith, editor; R. A. Heising, treasurer; F. B. Llewellyn, A. F. Van Dyck, H. A. Wheeler, and W. B. Cowilich, assistant secretary.

These applications were recommended to the Board of Directors for approval: for transfer to Member from H. E. Dinger, Beverly Dudley, T. M. Ferrill, Jr., J. B. Knight, Clarence Radius, A. G. Sparling, C. J. Terrey, and R. G. Zender, and for admission to Member from J. A. Cose, D. M. Davis, J. E. Innes-Crump, G. F. Maedel, O. S. Meixell, A. C. Omberg, Charles Sheer, and W. D. Wenger.

The 110 applications for Associate, 120 for Student, and 5 for Junior were also approved for confirming action by the Board of Directors.

Consideration was given to the proposals made by B. J. Thompson, chairman of the Membership Committee, regarding the current year's activities planned for this group.

It was decided to make certain temporary changes in the Institute's Code of Administrative Practice.

The bank and other resolutions were approved for submission to the Board of Directors.

It was noted that members in the Armed Services were maintaining active memberships and, in cases where they were unable to receive the PROCEEDINGS, were having their issues held in reserve by the Institute.

Chairman Wheeler reported on his recent letter to H. S. Osborne, president of the American Institute of Electrical Engineers, expressing appreciation of that society's co-operation received by the Institute during its Winter Conference in New York City.

The assistant secretary pointed out that the attendance at the Winter Conference on January 28, 1943, at New York City, was approximately 450 members and guests at the morning session, 350 at the afternoon session, and 600 at the joint evening meeting.

Chairman Wheeler stated that the Institute had sent flowers to the funeral of the late Dr. Nikola Tesla, held in December.

Approval was granted to paying the earned advertising commissions, requested by W. C. Copp, on the basis of twice a month. It was also authorized to accept the actual charge for making the recently completed audit of the Institute's 1942 financial records.

Treasurer Heising, in his capacity as chairman of the Constitution and Laws Committee, gave a brief oral report on the constitutional changes in process.

Editor Goldsmith explained the current restrictions on the use of paper in periodical publishing and its effect on the PROCEEDINGS, and made a paper-saving proposal relative to the PROCEEDINGS which was approved.

The existing Technical Committees were discussed and consideration was given to their personnel for the next annual term, which begins May 1, 1943.

Three members of the Executive Committee were appointed and each placed in charge of a division of the Institute's committee activities, as required by the Institute's Code of Administrative Practice.

Letters were read from C. L. Fortescue, president, and R. L. Smith-Rose, both of the Institution of Electrical Engineers, London, on the subject of co-operation between that society and the Institute. A discussion was also held on a previously prepared memorandum relative to the general subject of the Institute's foreign-membership policy. These matters were referred to the Board of Directors.

The recent letter from the American Association for the Advancement of Science, concerning the formulation of general objectives for science, was discussed and likewise included for consideration of the Board of Directors.

A meeting of the Executive Committee was held on February 3, 1943, and those in attendance were L. P. Wheeler, chairman; Alfred N. Goldsmith, editor; R. A. Heising, treasurer; Haraden Pratt, secretary; A. F. Van Dyck, and W. B. Cowilich, assistant secretary.

Approval was granted to salary increases of certain office employees on the basis of individual-merit increases within established salary-rate ranges, or the operation of an established plan of salary increases according to length of service.

A policy was adopted relative to overtime work.

Winter-Conference Section Meetings

The following accounts describe those Section. meetings on which reports have been received by the Institute Office and which were held on or near January 28, 1943, the Winter-Conference Day set by the Institute's Board of Directors

BOSTON

The Boston Section conference preceded by a dinner, took place on January 28, 1943, at Northeastern University, Boston. The program included the following papers:

"Résumé of Impedance Measurements," (illustrated), by I. G. Easton, General Radio Company.

"Radio Production for the Armed Forces," by Rear Admiral S. C. Hooper, U.S.N. (Paper Presented by Lieutenant Commander R. B. Meader U.S.N.)

"Radio Standards Go to War" by H. P. Westman, American Standards Association. (Paper Presented by P. K. McElroy, General Radio Company.)

By special arrangement at this meeting, Dr. Lee A. Du Bridge was formally presented the Institute's Fellowship Award, granted to him "for engineering and administrative leadership in the development and application of new radio techniques."

Despite one of the worst blizzards in the area, the conference was attended by 54 members and guests. J. M. Henry, past chairman, presided.

CINCINNATI

The Cincinnati Section conference was held on January 28, 1943, at the University of Cincinnati.

The following papers were given:

"Radio Production for the Armed Forces," by Rear Admiral S. C. Hooper, U.S.N. (Presented by Lieutenant Piersall, U.S.N.)

"Radio Standards Go to War," by H. P. Westman, American Standards Association. (Presented by H. G. Harding, Crosley Corporation.)

Vacation Pictures on Kodachrome Color Slides, shown by J. R. Duncan.

The conference was attended by 30 members and their guests. H. Lepple, chairman, presided.

CLEVELAND

The Cleveland Section meeting, preceded by a dinner, took place on January 27, 1943 at the Cleveland Engineering Society quarters, Cleveland.

The paper of the evening, "The Behavior of Dielectrics over Wide Ranges of Frequency and Temperature," was presented by Dr. R. F. Field, General Radio Company.

The annual election of officers was also held.

Approximately 25 attended the meeting which was presided over by F. C. Everett, vice chairman.

DALLAS-FORT WORTH

The Dallas-Fort Worth Section met on the evening of January 28, 1943, at the Adolphus Hotel, Dallas. The program included these papers:

"Pulse Circuits, Specialized Applications" (illustrated), by C. W. Tittle, Southern Methodist University Pre-Radar School.

"Square-Wave Analysis," by V. N. James, Southern Methodist University Pre-Radar School.

Major Strobart, U. S. Signal Corps, was called upon and spoke briefly of his interests in the Institute.

A. N. Stanton read an editorial from *Electronics* magazine concerning the warning of the attack on Pearl Harbor by the Radar operator on duty. The reading was followed by a discussion of the part that Radar is playing in the present war.

The radio addresses of incoming president L. P. Wheeler and retiring president A. F. Van Dyck, from New York, and that of Honorable James Lawrence Fly, from Washington, were received at 9:30 P.M. at the meeting.

The annual meeting and election of 1943 officers also took place.

Nearly 60 members and guests were in attendance. Chairman Truett Kimzey presided.

DETROIT

The conference meeting of the Detroit Section, on January 28, 1943, took place at the Engineering Society of Detroit quarters, and was preceded by a dinner.

The following papers were given:

"Electronic Devices in Industry," by R. A. Powers, Bundy Tubing Company.

"Radio Production for the Armed Forces," by Rear Admiral S. C. Hooper, U.S.N. (Presented by A. B. Buchanan, Detroit Edison Company.)

The radio addresses of incoming president L. P. Wheeler and retiring president A. F. Van Dyck, from New York, and that of Honorable James Lawrence Fly, from Washington, were heard over the local station WJR at 10:30 P.M. on a receiver installed at the meeting.

R. A. Powers, vice chairman, presided and 30 members were present.

MONTREAL

The Montreal Section held its conference on January 28, 1943, at McGill University, Montreal, the meeting being preceded by a dinner.

The program included the following papers:

"The Behavior of Dielectrics over Wide Ranges of Frequency and Temperature," by R. F. Field, General Radio Company. (Presented by J. M. Conroy, RCA-Victor Company.)

"Analysis of Performance of Loop Antennas," by Dr. F. S. Howes, McGill University. (Paper discussed by Messrs. E. A. Laport, R. R. Desaulniers, L. T. Bird, and A. Oxley.)

The addresses by the Institute's new president L. P. Wheeler and retiring president A. F. Van Dyck, from New York and by Honorable James Lawrence Fly, chairman of Federal Communications Commission, from Washington, were heard on public-address system over landline from New York.

Nearly 90 were in attendance at the meeting, at which L. T. Bird, vice chairman, presided.

ST. LOUIS

The conference meeting of the St. Louis Section took place on January 28, 1943, at the studio of radio station KMOX, St. Louis. A dinner was also held in conjunction with the meeting.

The paper of the evening, "Frequency Modulation and Emergency Communications in a Public Utility," was presented by B. B. Miller, Public Service Company of St. Louis, and was followed by a general discussion of the subject.

The radio addresses of the Institute's new president L. P. Wheeler and retiring president A. F. Van Dyck, from New York, were received through the facilities of station KMOX, in addition to the address of Honorable James Lawrence Fly, from Washington. The subject of the latter address was discussed at length by those present.

O. S. McDaniel, chairman, presided and more than 50 attended the meeting.

SAN FRANCISCO

The San Francisco conference meeting, on January 28, 1943, was held at the studio of radio station KQW, Palace Hotel, San Francisco, and was preceded by a dinner.

The program included the following papers:

"A Wide-Band Cathode-Ray Oscilloscope" (illustrated), by Dr. E. D. Cook, General Electric Company. (Presented by P. G. Caldwell, General Electric Company.)

"Radio Production for the Armed Forces," by Rear Admiral S. C. Hooper, U.S.N. (Presented by Mr. Cameron, Heintz and Kaufman, Ltd.)

The meeting was opened with a reading of the address of retiring president A. F. Van Dyck, which was submitted by the Institute headquarters to sections holding simultaneous winter conferences.

The radio addresses, broadcast that evening, of the Institute's new president L. P. Wheeler and retiring president Van Dyck and of Honorable James Lawrence Fly, were received at the meeting.

At the meeting there were 60 present and Dr. Karl Spangenberg, chairman, presided.

TWIN CITIES

The January 28, 1943 conference meeting of the Twin Cities Section took place at the studio of radio station WCCO, Minneapolis, and included a dinner.

These papers were given and on them a round-table discussion followed:

"Some Problems in Filter Design," by Ralph Allison, Audio Development Company.

"The Winona (Minnesota) Police Department Frequency-Modulation Radio Installation," by W. A. Haeussinger, Radio Station KBZB.

"Monitor Crosstalk Problem in Station WCCO's Master Control Room," by A. G. Peck, Radio Station WCCO.

The radio addresses of the Institute's new president L. P. Wheeler and retiring president A. F. Van Dyck, and that of Honorable James Lawrence Fly, were received at the meeting over station WCCO, beginning at 9:30 P.M.

Approximately 25 attended the meeting, which was presided over by Ralph Allison, chairman.

WASHINGTON

The conference and banquet of the Washington Section were held on January 28, 1943, at the Willard Hotel, Washington.

The following papers comprised the technical program:

"The Radio Enginner in War Industry," by F. H. McIntosh, Chief of Domestic and Foreign Radio Branch, Radio and Radar Division of War Production Board.

"The Radio Engineer in the Army," by Brigadier General F. E. Stoner, Director of Army Communications, U. S. Army.

"The Radio Engineer in the Navy," by Captain C. F. Holden, Director of Naval Communications, U. S. Navy.

"The Radio Engineer in Psychological Warfare," by R. C. Corderman, Assistant Chief of Bureau of Communication Facilities, Office of War Information.

At 10:30 P.M. Honorable James Lawrence Fly, chairman of the Board of War Communications and the Federal Communications Commission, delivered the radio address, "Radio Engineering in Wartime," which was broadcast from Washinton over the Columbia Broadcasting System stations and received at the banquet on a public-address system.

Preceding Chairman Fly's address were those of the Institute's retiring president A. F. Van Dyck and new president L. P.

Wheeler, as part of the same radio program, from New York.

The banquet was followed by a full program of entertainment, with Arthur Godfrey as master of ceremonies. Dancing after the radio broadcast concluded the evening's activities.

The committee responsible for the conference arrangements, included Captain E. M. Webster, C. M. Hunt, Lieutenant Commander G. C. Gross, and H. A. Burroughs.

There were 275 members and guests in attendance at the conference and banquet.

OTHER SECTION MEETINGS

Los Angeles

"Television, Present and Future," by H. R. Lubcke, Don Lee Broadcasting system, January 19, 1943.

Montreal

"Physics and the Radio Engineer," by Dr. W. H. Watson, McGill University, February 10, 1943.

Philadelphia

"Effect of Solar Activity on the Ionosphere and Radio Communications," by Dr. H. H. Wells, Carnegie Institute of Washington, February 3, 1943.

Pittsburgh

"The Behavior of Dielectrics over Wide Ranges of Frequency and Temperature" (illustrated), by R. F. Field, General Radio Company, January 11, 1943.

Naval U-H-F Engineering Training

The attention of the readers of the Proceedings is directed to the following announcement from the Navy Department of the United States:

"The rapid expansion in Navy surface vessels, submarines, and aircraft has created several hundred additional billets for officers trained in electrical engineering who are needed to serve in engineering work in connection with ultra-high frequency electronic apparatus, the Navy announced today.

"The Navy has issued a call for qualified engineers to fill the new vacancies. Technically qualified for this work are men who hold degrees in electrical engineering and have practiced in the field of engineering since their graduation; or men who have majored in physics, mathematics, or other branches of alternating-current circuits and electronics.

"Men commissioned for this electronic work are given the Navy's three-months' course in ultra-high frequencies at either Harvard University or Bowdoin College, followed by an additional three-months' laboratory course at Massachusetts Institute of Technology. Upon graduation, these officers are assigned to responsible engineering positions having to do with research, design, instruction or maintenance of the Navy's ultra-high frequency equipment.

"Qualified engineers are urged to apply for a commission at the nearest Office of Naval Officer Procurement. These offices are located in principal cities throughout the United States."

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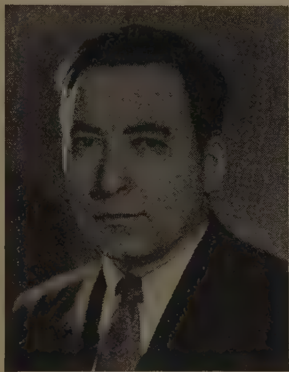


Brown Brothers

Ralph R. Batcher

In accepting a new post as consulting editor of *Electronic Industries*, recently inaugurated publication in the radio-and-electronic field, Ralph R. Batcher has added to his numerous and productive wartime activities. He is the author of textbooks and articles in the radio-and-electronic field, and has devoted particular attention to cathode-ray tubes and their associated circuits. During World War I Mr. Batcher was an instructor in radio theory at the Signal Corps School of the College of the City of New York and also at the old Marconi Institute. On his graduation from Iowa State College in 1920 he became an engineer with the Western Electric Laboratories where he stayed until 1924 when he became research engineer with A. H. Grebe and Company. During 1929–1930 he was vice president of the Decatur Manufacturing Company, resigning to do consulting work until 1938 when he became chief engineer of the Allen D. Cardwell Manufacturing Company.

Mr. Batcher joined the Institute as a Junior in 1916, transferred to Associate in 1918, and to Member grade in 1922. He is a member of its Board of Editors and also has served effectively on the Publications Committee of the Institute and the Committees on Symbols and Television. He is a member of Tau Beta Pi.

Hazeltine Electronics Corporation

The name of Hazeltine Service Corporation has been changed to Hazeltine Electronics Corporation, according to an announcement just made by W. A. MacDonald (A'19-M'26), president. This step follows completion of a program of plant expansion providing large additional facilities for electronics research and development.

"The tremendous growth in the use of electronic devices by the Army and Navy has greatly increased the responsibilities entrusted to the Hazeltine organization," said Mr. MacDonald. "Since the infancy of radio broadcasting, Hazeltine has been supplying new principles, circuits, techniques, and equipment, and Hazeltine developments today are playing an ever more vital part in helping to keep the United Nations superior to the enemy. With enlarged facilities in plant and personnel we can undertake solution of the most complex problems in electronics. Following the war, Hazeltine Electronics Corporation will be able to supply the public, through industry, with electronic marvels that can make life safer and happier for every one."

Bibliography to be Published on Automatic Stations

The fourth bibliography of technical literature entitled "Bibliography on Automatic Stations, 1930–1941" is soon to be issued by the American Institute of Electrical Engineers. This publication sponsored by the AIEE committee on automatic stations supplements earlier bibliographies on the subject published previously in *AIEE Transactions*.

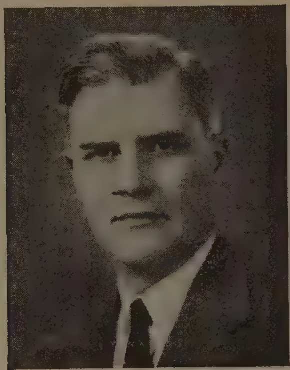
The entries in this bibliography are numbered consecutively by sections and listed alphabetically by years. The material is divided into the following sections:

General; supervisory and remote control; telemeters and telemetry; automatic and remote-controlled switches and switchgear; automatic features of generating stations using fuels; automatic boiler and combustion control; automatic hydroelectric plants; automatic substations.

The "Bibliography on Automatic Stations, 1930–1941" will be obtainable from AIEE headquarters, 33 West 39th Street, New York, N. Y., at 25 cents per copy for AIEE members (50 cents to nonmembers) with a discount of 20 per cent for quantities of 10 or more mailed at one time to one address. Remittances, payable in New York exchange, should accompany orders.

Radio Club of America Meeting

A meeting of The Radio Club of America was held on Thursday, February 11, at Columbia University, New York. The speaker was Robert F. Field, Engineer, General Radio Company, and his subject was "The Behavior of Dielectrics Over Wide Ranges of Frequency and Temperature." Mr. Field is a Member of the I.R.E. The summarizing paragraph of Mr. Field's paper indicates that he dealt with several interesting aspects of the behavior of dielectrics. The changes in dielectric constant and loss factor of several dielectrics were described. It was shown that two types of polarization, dipole and interfacial, are responsible for these changes. These polarizations are defined by three parameters whose values are found either from plots of loss factor against dielectric constant, or by a graphical analysis of the current-time curves of the dielectric. Several examples were given of the variation of the polarization parameters with frequency, illustrated by graphs.



John Kelly Johnson

John Kelly Johnson for many years engineer in charge of the Hazeltine laboratories in Chicago has resigned his position as senior engineer in Hazeltine Electronics Corporation in order to accept the position of Special Representative assigned to the Office of Procurement and Material of the Office of the Under Secretary of the Navy. He will assume his new duties with headquarters in Washington, D. C. immediately. Mr. Johnson joined The Institute of Radio Engineers as an Associate in 1925 and transferred to Member grade in 1935.

Dr. Howard A. Pidgeon

Dr. Howard A. Pidgeon, vacuum-tube development engineer of Bell Telephone Laboratories, died February 7 after a short illness. He was born in Pennsville, Ohio, on October 25, 1883.

Dr. Pidgeon received the B.S. degree from Ohio University in 1911 and the M.S. degree in 1912. He then went to Cornell University where he engaged in work in engineering and physics, acted as instructor in physics and received his Ph.D. degree in 1918. He joined the Laboratories, then the Engineering Department of the Western Electric Company, in October, 1918, and for a number of years was engaged in fundamental studies of electron emission. Since then, in the Electronics Research Department, he had been in charge of a group responsible for the development of all types of vacuum tubes used for communication purposes, particularly those used in modern long-distance telephone equipment. Since the beginning of the present war he had been concerned with the development of vacuum tubes and other electronic devices for the Army and Navy. Dr. Pidgeon had contributed several technical papers in his field of endeavor. He joined The Institute of Radio Engineers as a Member in 1929 and he was a member of the American Physical Society, New York Electrical Society, American Association for the Advancement of Science, and the National Geographic Society.

Irvin Ray Baker

Irvin Ray Baker, who was formerly head of RCA's Broadcast Transmitter Sales and latterly was advancing the development of electronic applications to war industries, died February 9, of a cerebral hemorrhage.

Mr. Baker was born on a farm in Freedom Township, Adams County, Pa., on October 6, 1903. He attended the public and high schools of the township and later entered Gettysburg College. He had become interested as a boy in what was then known only as wireless. After he received his B.S. degree, he continued his studies to receive another degree in electrical engineering.

Shortly after his graduation, Mr. Baker entered the employ of the General Electric Company in 1927. There he was placed in charge of operations of the Schenectady, New York, broadcasting station, WGY, one of the first in the nation. In 1929 he joined the Radio Corporation of America and in a few years he was appointed as head of broadcast transmitter sales at the RCA Camden plant. He was a Member of The Institute of Radio Engineers since 1935. When the war came, Mr. Baker devoted most of his time to research into the use of high-frequency radio current to speed war production.

Correspondence

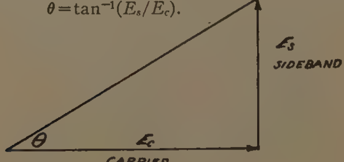
Correction of F-M Distortion

The frequency-modulation distortion-correction system described in Pieracci's article[1] has been the subject of much controversy. In this paper it was shown that the system was capable of correcting distortion for a sinusoidal applied signal. Many readers doubted the ability of the system to correct distortion when a complex signal was applied. This doubt was only partly dispelled by the supplementary article in the March, 1942, Proceedings.

The following is a mathematical analysis of the frequency-modulation distortion-correcting scheme. It includes a derivation of the necessary correction and a proof that if the correction is properly made for a sinusoidal input it is sufficient regardless of the nature of the applied signal.

In a frequency-modulation system, if the phase angle θ is made to vary linearly with the sideband E_s then there will be no amplitude distortion introduced at this point in the system. It is, therefore, desired that θ be related to E_s by the equation $\theta = KE_s$. From the triangle

$\theta = \tan^{-1}(E_s/E_c)$.



It is immediately obvious that E_c must be varied if the desired linear relationship is to be obtained. Assuming that it is possible to obtain the strict linear relationship,

$$\theta = KE_s = \tan^{-1}\frac{E_s}{E_c} \quad \text{or} \quad \frac{E_s}{E_c} = \tan KE_s.$$

From this last equation the desired form for E_c is obtained:

$$E_c = E_s \cot KE_s.$$

For small values of the argument the cotangent can be expressed by the series

$$\cot x = \frac{1}{x} - \frac{x}{3} - \frac{x^3}{45} - \frac{2x^5}{945} \cdots$$

so that

$$E_c = E_s\left[\frac{1}{KE_s} - \frac{KE_s}{3} - \frac{(KE_s)^3}{45} - \frac{2(KE_s)^5}{945} \cdots\right]$$

$$= \frac{1}{K} - \frac{KE_s^2}{3} - \frac{K^3E_s^4}{45} \cdots .$$

Taking now $E_s = 1$ for $\theta = 60$ degrees or 1.048 radians, $K = \theta/E_s = 1.048$. Hence, for complete correction E_c must have the form

$$E_c = 0.955 - 0.349E_s^2 - 0.025E_s^4 \cdots .$$

If now E_c is related to E_s in such a way that for a sine wave ($\sin pt$) applied (taking the numerical values given in the original article)

$$E_c = 0.765 + 0.188 \cos 2pt$$

then

$$E_c = 0.765 + 0.188(1 - 2 \sin^2 pt)$$

$$= 0.953 - 0.376 \sin^2 pt$$

or

$$E_c = 0.953 - 0.376E_s^2.$$

Hence, if the correction is properly made for a simple sine-wave input, then necessarily the carrier is related in the proper manner to the sideband voltage E_s to give a good approximation to a linear relationship between θ and E_s. (The high coefficient of E_s^2 here compensates somewhat for the neglected higher-order terms.) This being true the correction is properly made to compensate for any input wave regardless of how complex it may be.

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[1] Roger J. Pieracci, "A stabilized frequency-modulation system," Proc. I.R.E., vol. 30, pp. 76–81; February, 1942; Supplement, March, 1942, p. 151.

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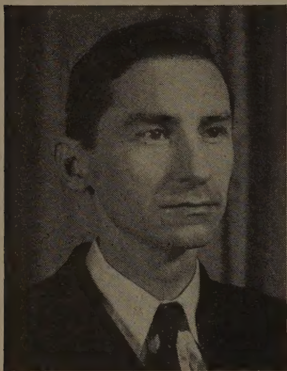
* Also Chairman of its Subcommittee on Insulating Material Specifications for the Military Services.

Contributors



Otto H. Schade

Otto H. Schade (M'40) was born on April 27, 1903, at Schmalkalden, Germany. He was graduated from the Reform-Real-Gymnasium, Halle, Germany, in 1922. From 1922 to 1924 he was with the Telephonfabrik A.G. vorm. J. Berliner, Berlin and Düsseldorf; from 1924 to 1925, in charge of the laboratory in the radio manufacturing company "Ratag" in Berlin; and from 1926 to 1931, in the engineering department of the Atwater Kent Manufacturing Company. He received the Modern Pioneer Award, February, 1940. Since 1931 Mr. Schade has been in the research and engineering department of the RCA-Victor Division, Radio Corporation of America, Harrison, N. J.

❖

C. M. Sinnett was born at East Livermore, Maine, in 1900. He received the B.S degree in electrical engineering from the University of Maine in 1924. Mr. Sinnett was in the radio engineering department of the Westinghouse Electric and Manufacturing Company from 1925–1930; in charge of the phonograph development and design department of the RCA Manufacturing Company from 1930 to 1933; on various special assignments in the engineering department of the RCA Manufacturing Company from 1933 to 1940; and in the advanced development division of the latter company on phonograph development from 1940 to date. He is a member of Tau Beta Pi.

C. M. Sinnett

❖

Frederick Emmons Terman (A'25–F'37) was born on June 7, 1900, at English, Indiana. He received the A.B. degree in 1920 and the degree of Engineer in 1922 from Stanford University, and the Sc.D. degree from Massachusetts Institute of Technology in 1924. From 1925 to 1937 Dr. Terman was an instructor, assistant professor, and associate professor of electrical engineering at Stanford University. Since 1937 he has been professor and head of the electrical engineering department at Stanford. In February, 1941, Dr. Terman was given a leave of absence by Stanford to assume the directorship of the Radio Research Laboratory which operates under Harvard University, Cambridge, Massachusetts, in co-operation with the Office of Scientific Research and Development. At the present time he is also a member of Divisions 14 and 15 of OSRD. Dr. Terman was Vice President of the Institute of Radio Engineers in 1940 and President in 1941.

Frederick E. Terman

❖

H. W. Wells (A'36–M'36) received the B.S. degree in electrical engineering in 1928 and the E.E. degree in 1937 from the University of Maryland. He was subsequently employed by Westinghouse Electric and Manufacturing Company at Pittsburgh, Pennsylvania. The All American Malaysian Expedition selected Mr. Wells in 1929 to take charge of radio communications for their expedition into Central Borneo. After completing the explorations in lands of the Dyak head-hunters, he joined the radio engineering staff of Heintz and Kaufman in 1930 engaging in receiver and transmitter development together with standard-frequency installations for Globe Wireless. This was followed by an interlude during which Mr. Wells joined the Army Air Forces, being graduated from Brooks and Kelly Fields in 1932, and served on active duty with pursuit squadrons at Langley Field. He has been a member of the scientific staff of the Department of Terrestrial Magnetism, Carnegie Institution of Washington, since late 1932, engaged in ionospheric and associated research in this country and in Peru, South America. During this period he has been instrumental in developing apparatus for multifrequency ionospheric exploration, in extending our knowledge of world-wide ionospheric characteristics, and in applying this information to theoretical analyses of the earth's magnetic field together with its practical application to radio wave-propagation. A professional degree, E.E., was received from University of Maryland in 1937.

H. W. Wells

❖

For a biographical sketch of G. L. Beers, see the Proceedings for March, 1943; for Arthur Van Dyck, see the Proceedings for January, 1943.